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ELEKTOR ELECTRONICS

video digitizer

Temperature-insensitive voltage divider

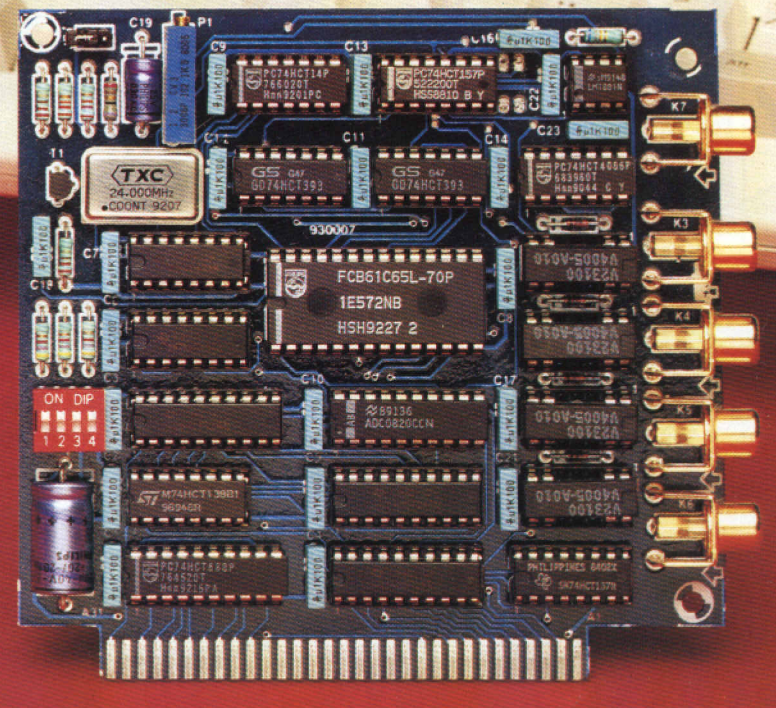
IR receiver for 80C32 SBC

Audio power meter

Wideband wireless data systems

27 MHz transmitter

Open-baffle loudspeaker system



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ELEKTOR ELECTRONICS

April 1993
Volume 19
Number 210

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- FM stereo signal generator
- Making sense of measurements
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- Philips preamplifier
- External ferrite antenna for SW, MW and LW radios
- RDS decoder
- EMC - Part 1

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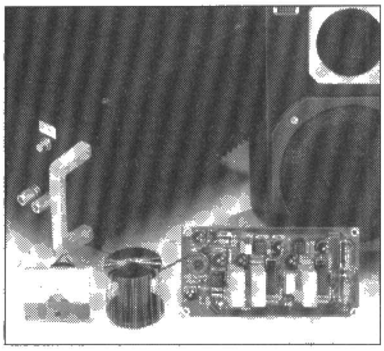
Front cover

The video digitizer insertion card (described on page 32) allows you to put TV images on to your computer screen. Designed to handle colour images at a resolution of 24 bits, the card forms a perfect link with several graphics workshop programs. The software supplied with the card packs the digital colour information into a TIFF (Tagged Image File Format) file, which can be processed further by almost any graphics program that runs under MS-DOS or Windows.



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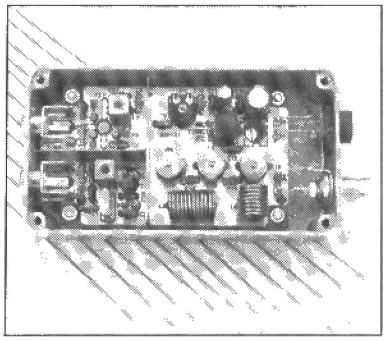

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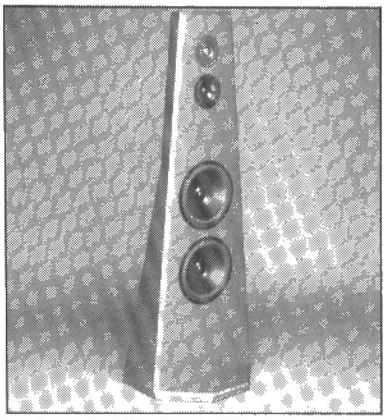

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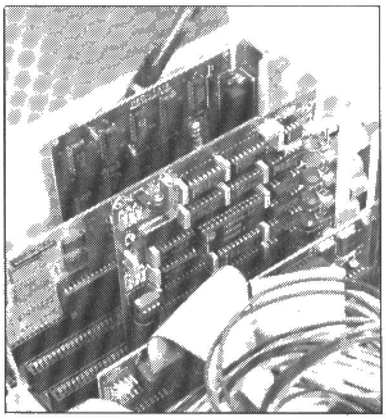

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ELECTRONICS SCENE

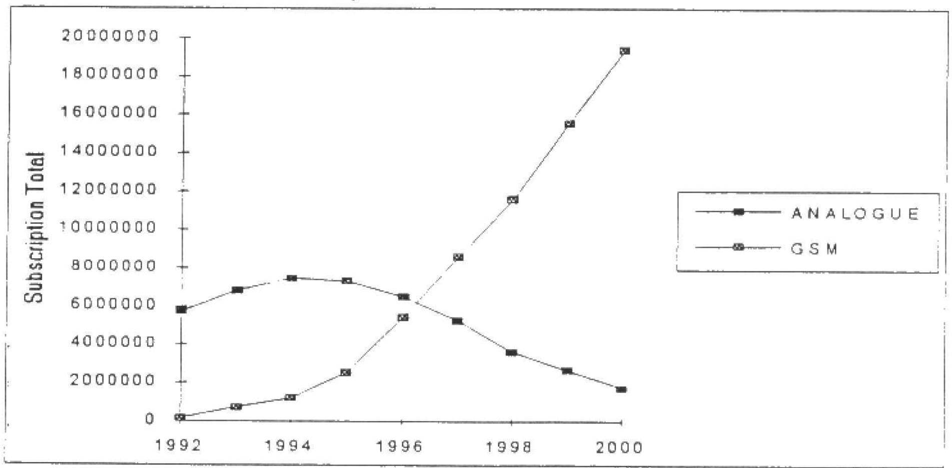
RADIO-TELEPHONE SYSTEMS IN EUROPE

By the year 2000, the European market for cellular mobile radio-telephone systems will total 19 million GSM (Global System for Mobile communications) subscribers, says InfoCorp's latest industry report. If oprators are successful in their bids to penetrate the mass market, subscriber levels could reach 23 million. According to InfoCorp, the reason for this rapid growth is that competition, which until now has been sporadic in this sector, will stimulate the market (as can be seen in the UK—Ed). Each European country has or will have at least two GSM operators. Moreover, terminal manufacturers will be able to benefit from economies of scale as a result of a single standard implementation. Lower prices for cellular phones and implementation of more effective distribution channels will result in a substantial increase in the number of GSM sunscribers.

According to the report, the size and evolution of the GSM market will be closely related to the cost of service to the end-user. Cellular market penetration levels are highly variable (from 0 to 77 subscribers per 1000 inhabitants). Countries with high market penetration like the UK and Sweden have the lowest tariffs. In other countries, GSM's market potential will be realized only in the medium and long term. In the short term, testing the system will be necessary and could hinder GSM subscriber growth.

InfoCorp Europe, 12 Boulevard des Iles, 92441 Issy-les-Moulineaux, France.

European GSM and Analogue Subscriber Totals in the Year 2000:



Source: InfoCorp Europe

MAPPING THE WAY FOR RADIO WAVES

Satellite pictures of Earth are being used to improve communications and broadcasting by ensuring that transmitters are located in areas that offer the best spread of radio waves.

Although Britain is believed to be the best mapped country in the world, surprisingly little is known about the type of ground cover—that is, to what extent the areas are built up or forested—that affects the spread of radio waves. Such information is vital to enable scientists to create computer models of radio wave propagation and attenuation. Along with information concerning radio system performance, it enables the interaction between radio waves and the earth's surface to be accurately reproduced in such models.

Deriving clutter information from mapping and aerial photography, even where sources are avalaible, is both time consuming and costly. In addition, for radio organizations, map information is often difficult to relate to useful ground clutter specification.

To overcome these problems, Britain's National Remote Sensing Centre (NRSC) has carried out a pilot project designed to privide digital terrain cover information of particular countries or areas in a cost-effective and easy-to-use way.

Its scientists have used satellite images to classify ground cover on a regional basis as water, wetland, agricultural land, semi-natural vegetation, non-agricultural bare land, deciduous or coniferous woodland, urban or dense urban areas. Within large urban areas, a further breakdown into dense, industrial, suburban areas and open suburban spaces was also carried out, and transport routes were placed in a variety of classifications.

The project, commissioned by a consortium of UK government and industrial organizations, enabled NRSC to provide a land cover database of a 185 km^2 area around London and the Thames estuary, derived from imagery acquired by the Landsat and SPOT satellites.

By using existing maps in conjunction with fieldwork, researchers were able to establish the appearance characteristic of different cover types in the satellite images, and to locate landscape features in the images to an accuracy of 100 metres.

ELIMINATING CFCs FROM AN ELECTRONIC MANUFACTURING PROCESS

Chlorofluorocarbons (CFCs), when emitted into the atmosphere, have been found to deplete the stratospheric ozone layer that shields the earth from the sun's harmful ultraviolet rays. The CFC layer in the atmosphere also reflects heat emissions back to the planet and, with halons, are also believed to make a significant contribution to global warming. Though accounting for only a small proportion of the greenhouse gases, they are 10 000 times more powerful than carbon dioxide and could account for up to 20% of the projected warming.

The International Montreal Protocol calls for the elimination of CFCs and other ozone-depleting substances by the year 2000.

It is not easy to find substitutes for CFCs in all its applications, and the electronics industry has the most problems.

CFC usage in the electronics industry is mostly in cleaning with CFC-113 (CCl_2F-$CClF_2$). Britain's Northern Telecom (NT) pledged in 1988 to eliminate CFC-113 from its manufacturing operations in three years. This goal was achieved nine years ahead of the mandate set by the Montreal Protocol.

Water-soluble pastes. The achievement of becoming the world's first major electronics company to eliminate ozone-depleting CFCs from its operations owes a great deal to work at BNR Europe in Harlow, near London, begun by what was then STC Technology (STL), a leading British telecommunications firm. Indeed, a programme was already under way at STL Harlow to find alternatives to CFC solvents in cleaning flux residues from PCBs and hybrids after assembly.

Initial work, started in 1988, was on water washing processes for the group's International Computers Ltd (ICL). As a result, ICL's Kidgrove plant moved completely to water-soluble pastes and fluxes in mid-1990. This application on a high-volume production line brought cost reductions and a higher yield while avoiding the use of CFC solvents.

Unfortunately, water washing is not suitable for all types of application and

components and, after some early work in 1989, a collaborative programme began in 1990 with GEC and British Aerospace. Partly funded by the Department of Trade and Industry and the Ministry of Defence, the programme has assessed all the options to using CFC-113 solvents. Alternative solvents and various solder fluxes and pastes have been evaluated.

However, says NT, the major achievement has been the feasibility of completely eliminating the need for solvent cleaning by using an inert atmosphere during soldering. This allows the use of less active fluxes, which do not need to be cleaned off after assembly. Using an oxygen-free atmosphere has also meant less contamination of the final product.

Improved reliability. In March 1991, NT integrated STC into its European operations to form Northern Telecom Europe Ltd, with headquarters in London, and sales and marketing operations across Europe, the Middle East, Africa and India.

Further NT-funded work is showing that not only can a clean PCB be produced at the same cost as before, but that the reliability of the product has been improved while eliminating CFC pollution. Work is continuing that will assure customers of the suitability of the process for products with even the highest reliability needs.

Inert atmosphere technology has been implemented at Cwmcarn in Wales, and Monkstown in Northern Ireland, while other sites are being helped with solutions according to their individual needs.

Northern Telecom estimates that the alternatives it has developed will prevent nearly 9 000 tonnes of CFCs being released into the atmosphere by its plants between now and the year 2000. The new processes should also save more than $50 million during that time in CFC purchasing costs and other expenses.

The company's achievement has drawn praise from the British Parliament and it received the 1992 North American Environmental Leaders award for outstanding environmental achievement from the United Nations Environment Programme (UNEP). The United States Environmental Protection Agency (EPA) estimates that 11% of the world's CFC emissions come from electronics manufacturing processes similar to the ones NT has successfully changed. Eliminating CFC solvents from these processes throughout this industry would have significant environmental effects.

PROGRESS TOWARDS TERRESTRIAL DIGITAL HDTV

Engineers believe it will soon be possible to transmit two high definition television (HDTV) programmes through one standard TV channel.

The engineering department of the British Broadcasting Corporation (BBC) says the use of a novel high spectral-efficiency modulation technique, together with advances in digital compression, have made it possible in principle to transmit HDTV programmes in a single 8 MHz channel.

BBC Engineering and France's Thomson-CSF Laboratories Electronique de Rennes (TSCF/LER) organization have collaborated in a successful experimental digital transmission using high spectral-efficiency modulation as part of a major project to develop the technology to bring digital HDTV to the general public via terrestrial transmission networks.

A digital TV signal was broadcast in a standard 8 MHz television channel from a low-power transmitter at the BBC's Crystal Palace mast in London. The signals were successfully received at the BBC's engineering research department and at a number of other test sites in south London and the adjoining county of Surrey.

The particular modulation technique used and the transmitting and receiving equipment were developed to convey about 60 Mbit/s in a single 8 MHz UHF television channel. The achievement of a spectral efficiency of about 7.5 bit/s/Hz is seen by engineers as a major development.

The system transmits two separate 30 Mbit/s signals, one of which is broadcast with horizontal polarization, the other with vertical polarization. Each 30 Mbit/s signal comprises an orthogonal frequency division modulation (OFDM) ensemble of about 500 closely spaced carriers, all of which are digitally modulated using 64-QAM.

In addition to this work, the BBC and Thomson are collaborating with other partners from several European countries in the development of a standard for digital terrestrial television broadcasting.

BBC Engineering Information, White City, 201 Wood Lane, London W12 7TS.

EUROPE'S NEW AIRBORNE TELEPHONE SYSTEM FLIES

A British digital terrestrial flight telephone system (TFTS) will offer European airline passengers a wide range of on-board telecommunication services, including in-flight telephone calls, facsimile and data transmissions worldwide.

Developed by GEC Sensors from Basildon, the TFTS provides a direct air-to-ground radio link for communication with a dedicated network of ground stations, which in turn interface with the international public switched telephone

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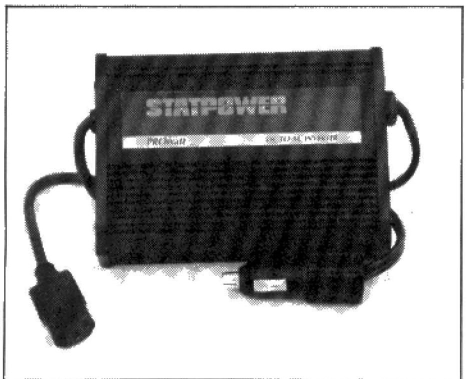
system. It conforms to the new performance standard defined by the European Telecommunications Standards Institute (ETSI), ensuring that aircraft can operate throughout Europe using a single-design standard of airborne equipment.

MAINS POWER IN YOUR CAR



The Merlin Pro Watt range of DC–AC inverters can be plugged into the cigarette lighter in your car or wired direct to the battery. The unit simply and silently converts 12 V d.c. into 240 V a.c. There are four models with outputs of 125 W, 200 W, 800 W and 1500 W. **Merlin Equipment, Scotts House, Cuxham Road, Watlington OX9 5JW, England.**

BT URGES MORE EUROPEAN LIBERALIZATION

BT has urged the European Commission to introduce full liberalization of telecommunications networks and services in the European Community by 1995 to comply with the requirements of the Treaty of Rome.

BT notes that the Community lags significantly behind its main trading partners, particularly the USA and Japan, in telecommunications liberalization, making it vital to take urgent steps immediately. BT said that it was for those opposing competition to justify why it should not prevail in telecommunications as it does in almost all other sectors. The competitiveness of Europe's industry depends heavily on its telecommunications networks and services.

Experience in the UK, USA and Japan has demonstrated that competition and universal service are compatible, so concern on those grounds is unfounded. Because the Community has fallen significantly behind, a Directive should be issued urgently requiring the withdrawal of special and exclusive rights in respect of all public voice telephony services and infrastructure.

WORLD'S FIRST DIGITAL RADIO BROADCASTING L-BAND SINGLE FREQUENCY NETWORK

Last December, the Canadian Broadcasting Corporation (CBC) Engineering, in collaboration with the Department of Communications (DOC), the Canadian Association of Broadcasters (CAB), the Communication Research Centre (CRC) and the Centre commun d'études de télédiffusion et télécommunications (CCETT), completed the implementation of an experimental Digital Radio Broadcasting (DRB) L-Band two-transmitter network (SFN). The projct was carried out under the auspices of the Joint Technical Committee on Advanced Broadcasting, and is part of a field test programme to characterize the L-Band frequencies for terrestrial DRB as well as to experiment with various transmitter/network configurations.

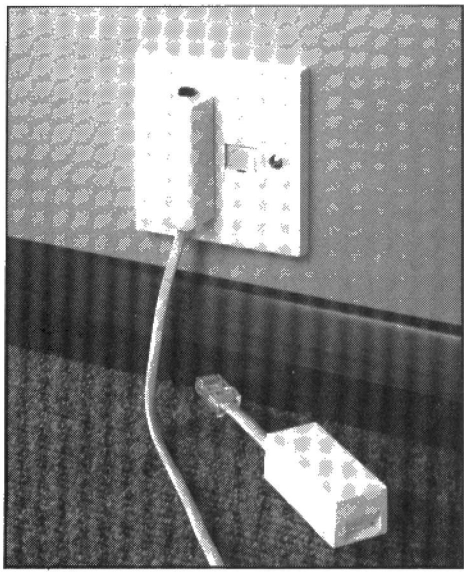
The facility consists of two transmitters broadcasting at the frequency of 1468.75 MHz in the band recently allocated to terrestrial and satellite Digital Audio Broadcasting (DAB) at the World Administrative Radio Conference (WARC-92). The first transmitter is located at the CN tower in Toronto and is operating with an output power of about 200 W. The transmit antenna is at a height of 364 m above ground and its gain is 17.85 dBd. The resultant ERP is about 9 kW. The antenna beamwidth is 60° and its orientation is towards the second site, which is located in Barrie at CKVR-TV site. The second station is transmitting 150 W in a 40° beamwidth antenna with a gain of 20.35 dBd, producing an ERP of around 16 kW. The transmit height is 232 m above ground. The two sites, which are 82 km apart, are transmitting the same signal on the same frequency synchronously. A microwave link is used to feed the Barrie site with the same signal as Toronto. A digital delay is used to compensate for the transmission delay.

On the basis of detailed propagation predictions and COFDM coverage simulations, it is expected that this facility will provide qualified (90–99%) continuous coverage in the area between the two transmitters in a width described by the antenna patterns of the sites. In theory, with two omnidirectional patterns, it would be possible to cover an area of about 40×160 km with two transmitters of ERPs in the 10 kW range, using the same frequency, at L-Band with a DAB system that can use on-channel re-transmitters, such as the COFDM system.

SLIMLINE RJ45 TO UK TELEPHONE ADAPTOR

A new adaptor from ITT Cannon provides a simple means of connecting a telephone into a data network cabling system. Measuring 35×56×24 mm, the adaptor is small enough to fit into a shallow floor box, while two units will plug into a standard dual UK wall outlet. A fly-lead version is also available with 150 mm of cord; other lengths can be supplied on request.

The adaptor incorporates a shuttered

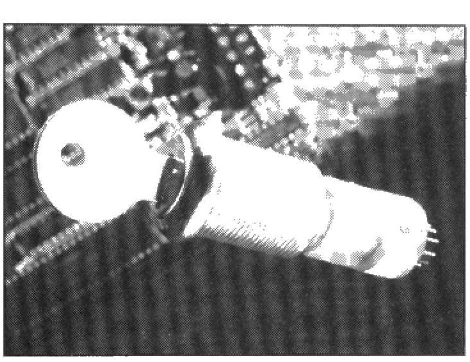


UK telephone socket, which is supplied with a security latch mechanism for the RJ45 to prevent unauthorized removal from an outlet.

The adaptors comply with Oftel approval NS/G/23/L 100005 and BS 6312. **ITT Cannon, Jays Close, Viables Estate, Basingstoke RG22 4BW.**

NEW KEYLOCK SWITCH

Grayhill's 58J economy rotary keylock is available in two configurations: as a minimum-space type with a depth of only 25 mm (1 in); or as an antistatic type providing 15 kV d.c. static protection.



The lock includes five tumbler plates for added security and is available with standard or spring return an in-panel key re-coding.

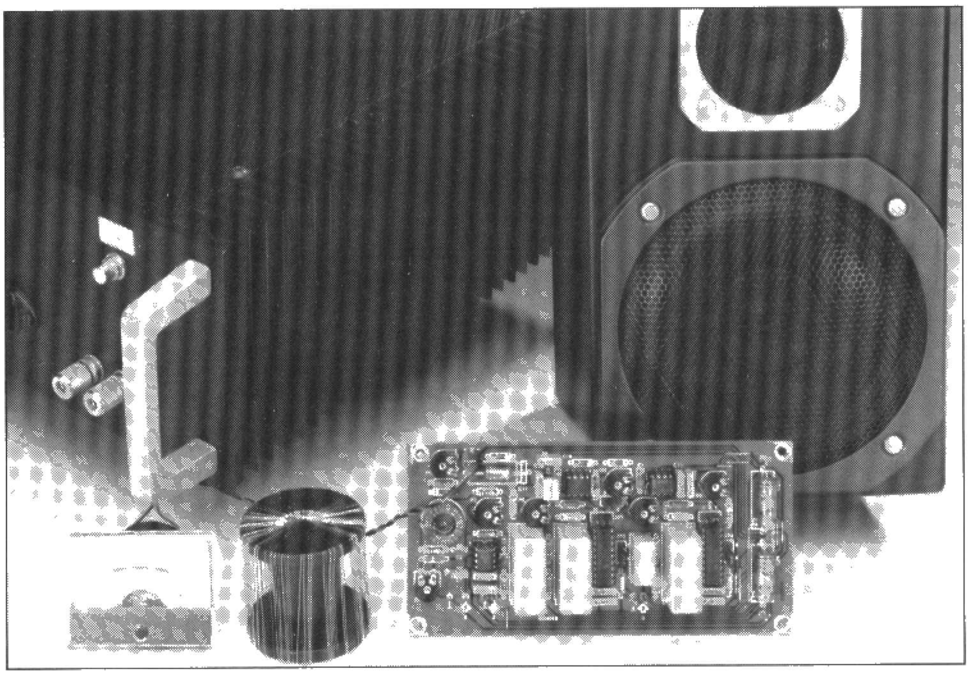
For switching applications, solder-lug or PCB mounting is available. Each switch can have one or two poles with up to ten positions per pole and throws of 36°, 45°, 60° or 90°.

The switch is available from **Highland Electronics Ltd, Albert Drive, Burgess Hill, RH15 9TN, England.**

AUDIO POWER METER

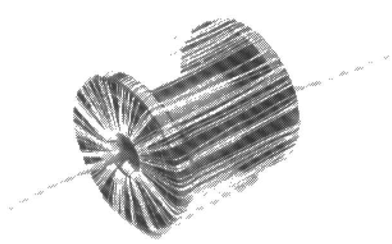
Design by T. Giesberts

Most audio output measurements are concerned with the voltage delivered across the loudspeaker by the output stage. If the loudspeaker impedance were known, it would be easy to calculate the power. But the impedance is known only approximately and, moreover, varies with frequency. Measuring both the current through, and the voltage across, the loudspeaker voice coil is, therefore, a far more reliable method of determining the power delivered to the loudspeaker.



Most measurements on output amplifiers concern voltage, which is not surprising, because current measurements invariably mean connecting a device in series with the loudspeaker and that degrades the damping of spurious cone movements. The present design is no exception: a device, here a current transformer, is connected in series with the loudspeaker. However, a current transformer affects the performance of the output stage much less than a soldered connection.

The current transformer used in the present design is an air-cored toroidal type whose construction is shown below.



The primary 'winding', running through the centre of the toroidal secondary winding, is nothing more than a wire through which the loudspeaker current flows. The toroidal shape has several advantages: (1) it collects most of the magnetic field around the (primary) wire; and (2) it almost totally rejects stray fields. Moreover, since it is air-cored, it is not affected by frequency. Strictly speaking, it is not a true current transformer, but rather a current-to-voltage transformer. Both theory and practice show that the output voltage of such a transformer is directly proportional to the current through the primary wire and to the frequency of the current flowing in that wire—see box on page 10.

A little artihmetic

A power meter must have the facility of multiplying current and voltage. In analogue meters this is achieved by the special construction of the meter. However,

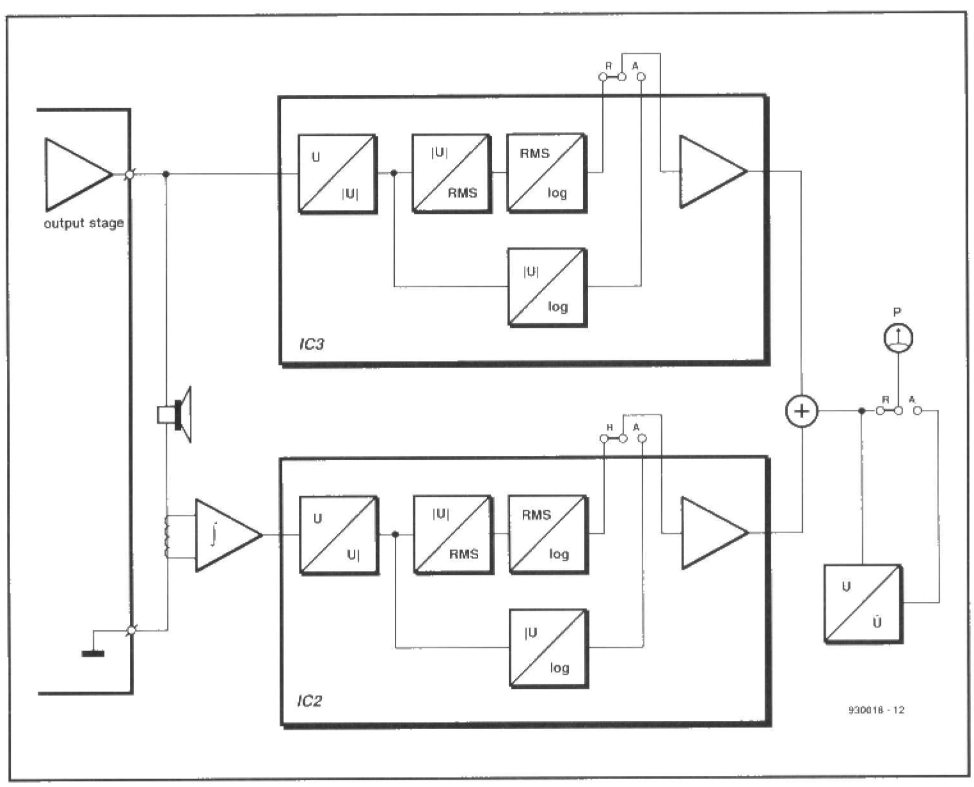


Fig. 1. Block diagram of the audio power meter.

multiplying in analogue electronics is a little tricky. Fortunately, the multiplication can be circumvented by using the fact that most power meters have a logarithmic scale. If this were not so, the meter would deflect only at maximum output power. Our hearing also reacts logarithmically to sound energy. Since the sound pressure is directly proportional to the delivered power, it is common sense to use a logarithmic scale. The instrument covers a measuring range of 100 μW to 100 W, enabling it to cope with large and very small signals alike.

The meter indicates log*P*, that is, log(*UI*), which may written as log*U*+log*I*. If only discrete components were used, the latter would not be any easier to compute than the former. Since, however, the logarithms of both *U* and *I* are computed by ICs, the calculation becomes simple.

The current through, and the voltage across, the loudspeaker are used as input signals to the power meter—see Fig. 1. The current is converted into a voltage by the transformer. As already mentioned, that voltage is directly proportional not only to the current, but also to the frequency. That relation to frequency is negated by an integrator, whose output is a signal that is directly proportional only to the current through the voice coil. Both inputs are taken to RMS/log converters. These devices have two outputs: (1) the logarithm of the RMS input, and (2) the logarithm of the rectified, absolute input signal (|*U*|), which enable the apparent power, *S*, or the instantaneous active power, *p*, to be measured. The apparent power is simply the product of voltage and current, irrespective of the phase difference between them. The active power is the product of instantaneous current, instantaneous voltage and the cosine of the phase angle, ϕ, between them.

When apparent power is measured, the logarithms of the RMS values are simply added and displayed on a meter. Note, however, that, since use is made of an *RC* network, the output signal of the converters is delayed slightly. It is, therefore, not possible to measure peak values in this mode. This is, however, possible when the absolute value of the input signals is used. The meter is then connected to a peak detector via a jump lead. The instantaneous active power is the power delivered to the loudspeaker by the amplifier at any given instant.

Effect of metering

Nothing has been said yet about the effect the audio power meter has on the amplifier/loudspeaker system. It is essential that the resistance of the link between amplifier and loudspeaker is as small as possible. In the circuit diagram in Fig. 2, the secondary winding of the current transformer is terminated by a resistance of not more than 150 Ω (R_1 and P_1). That resistance is reflected in the loudspeaker connection. If the primary winding of transformer L_1 is taken as 1 turn

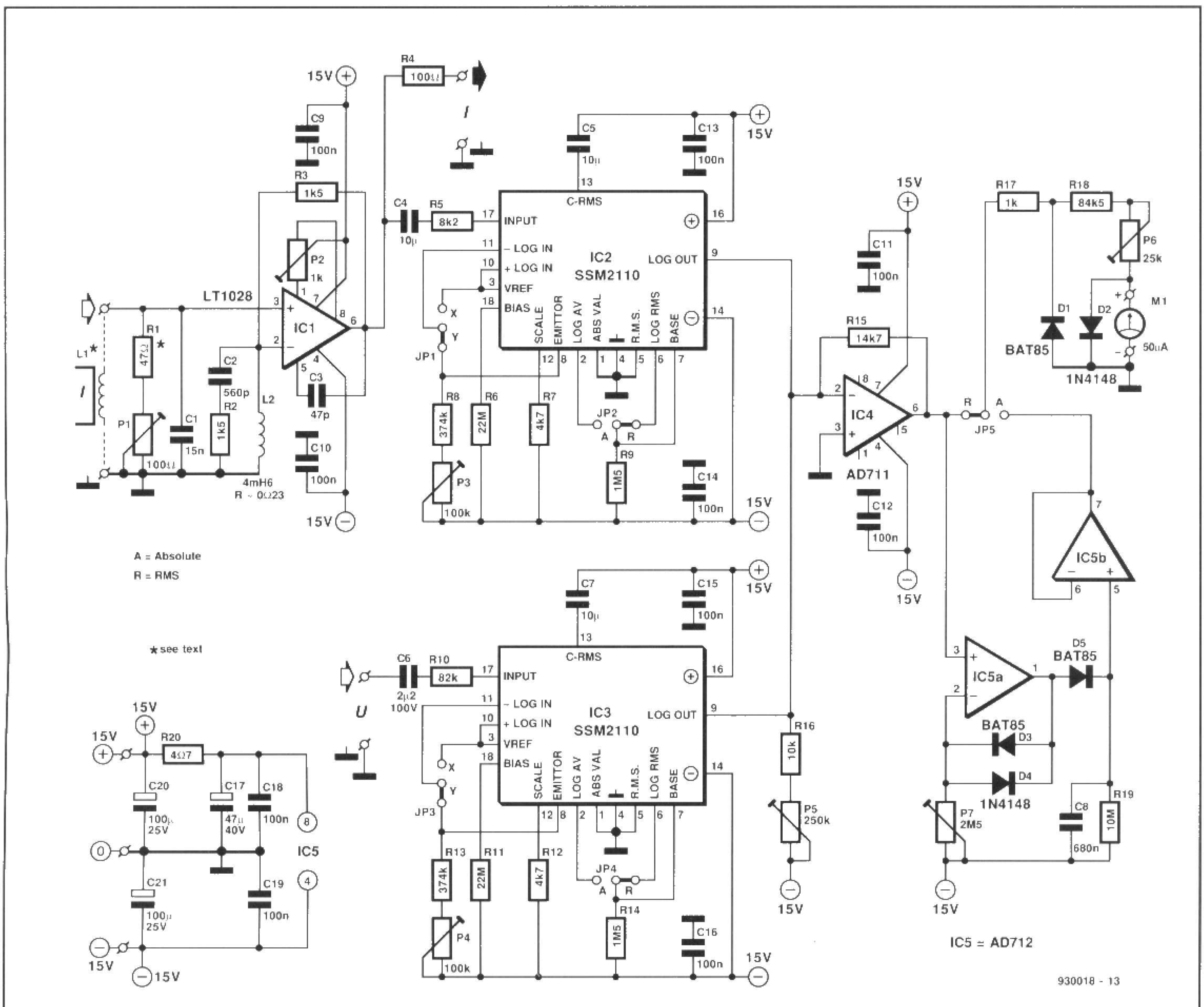


Fig. 2. Circuit diagram of the audio power meter.

(in reality not even half a turn), and the secondary winding consists of 150 turns, the reflected (additional) resistance in the loudspeaker connection amounts to $(^1/_{150})^2$ Ω, or just under 7 mΩ, which is appreciably less than that of a solder joint. If this is considered too high, the number of turns may be increased and the dimensions of the transformer reduced: this will be reverted to later.

To minimize the effect of parasitic capacitances between the transformer and its surroundings, it is essential that the transformer is located in the earth line to the loudspeaker—see Fig. 1.

Resistor R_1 and preset P_1 form not only a terminating resistance, but, together with L_1, are also constituent parts of the integrator based on IC_1. This is the reason that their values cannot be much lower than shown.

The integrator is a rather unusual design. It does not invert and its feedback loop does not contain a capacitor but an inductance, L_2. The reason for an inductance is that at 20 Hz the amplification is still fairly high (about ×2500). If a capacitor were used, the resistance between earth and the –ve input of the integrator would quickly rise to 1 kΩ and higher. Combined with the (relatively) large amplification factor, this would result in an appreciable increase in noise (the larger the impedance, the larger the noise it causes). On the other hand, a good inductance has little resistance and thus causes little noise. Its combination with the low-noise opamp ensures that the output signal is substantially free of noise.

If the integrator were an inverting type, L_1 and L_2 would be in series, which would result in the transformer affecting its operation. With a non-inverting integrator that is hardly the case. Hardly, because a non-inverting integrator (like a non-inverting opamp) has a minimum amplification of ×1. With the design shown in Fig. 2, IC_1 gradually ceases to function as an integrator from about 20 kHz and stops altogether at about 50 kHz. This process is compensated by low-pass filter L_1-R_1-P_1, which has a cut-off frequency of around 50 kHz (presettable with P_1). In other words, the gradually declining integration process of IC_1 is compensated by the gradually increasing integration provided by the filter. This means that, via a backdoor, L_1 influences the integrator after all.

Capacitor C_3 provides some frequency compensation of the opamp to ensure that this remains stable even at an amplification of ×1. To make sure that at frequencies greater than about 190 kHz this stability is retained, the amplification is increased to ×2 by C_2 and R_2. At the same time, C_3 ensures that such high frequencies do not even reach the input of the opamp. The offset of the opamp is compensated with the aid of preset P_2.

The output signal of the integrator and the voltage across the loudspeaker are applied to RMS-to-LOG converters IC_2 and IC_3 respectively. Jumpers JP_1 and JP_3 were used for experimental purposes, but have no longer any significance. They are replaced on the PCB by wire bridges Y. Jumpers JP_2 and JP_4 enable either the logarithm of the RMS value or that of the absolute value of the signal to be selected. From these jumpers, the signal is fed back to a buffer amplifier in the relevant IC. Presets P_3 and P_4 provide fine adjustment of the converters, which in most cases will not be necessary.

The signals are output by the buffer amplifiers as currents, so that they can be added together and applied to the input of counter IC_4. An offset is added to the sum signal via preset P_5 to set the lower end of the measuring range correctly to 0 V.

The output of IC_4 is applied to peak detector IC_{5a} and to jumper JP_5. With this jumper in position R, the output of IC_4 is also applied to the meter resistors. Diode D_1 ensures that no large negative potential can arise across meter M_1, while D_2 prevents too high a voltage across the meter. Full-scale deflection of the meter is set with P_6.

The negative feedback of peak detector IC_{5a} is designed to give the opamp unity amplification. The offset current set with P_7 causes D_3 and D_4 to conduct to an extent that results in a potential of 15 V across R_{19}-C_8 when the minimum power that the meter can display is present at the input of the detector (the potential at junction R_{19}-C_8-D_5 is then 0 V). In that way, compensation is obtained for the voltage drop across D_5 and this ensures that the true peak value of the input signal is measured. Since C_8 is charged fast (only the internal resistances of IC_{5a} and D_5 limit the charging current), but discharges only slowly via R_{19} (time constant R_{19}-C_8 = 6.8 s), the peak value is retained long enough for M_1, a moving-coil types that takes a few seconds to deflect, reading the correct value. To prevent the meter loading C_8, buffer IC_{5b} separates them.

Construction

The audio power meter is best constructed on the printed-circuit board shown in

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The size of the transformer for a given output amplifier may be calculated from the following formulas.

$H = I/2\pi r$;

$B = \mu H$;

$\phi = BA$;

$U = \mathrm{d}\phi/\mathrm{d}t$,

ℓ = length in metres (m)

where H is the magnetic field strength at distance r from the conductor that carries current I; μ is the magnetic permeability (here, of air, which is $\mu_0 = 4\pi \cdot 10^{-7}$); ϕ the magnetic flux through the inductor; A the profile area of the inductor, ℓb, and U the voltage measured across the coil.

The voltage, U, across a coil of N turns resulting from a current I through the coil is given by

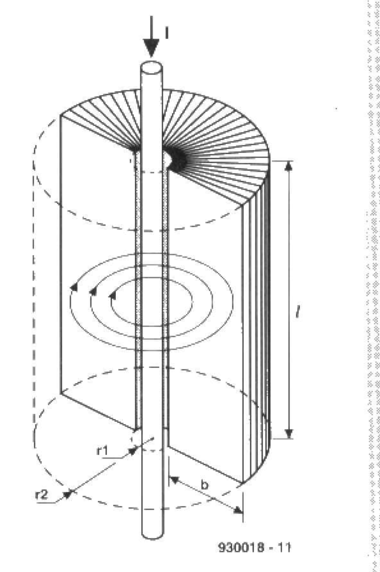
$$U = N\mathrm{d}(\mu_0 H \ell b)/\mathrm{d}t.$$

The average fieldstrength, H, between r_1 and r_2 is calculated from

$$\int_{r_1}^{r_2} H \cdot \mathrm{d}r = \int_{r_1}^{r_2} [I/2\pi r]\mathrm{d}r$$

$$= [I/2\pi] \cdot \log_e[r_2/r_1].$$

This result is not just the average field strength between r_1 and r_2, but also the field strength times width, b $(=r_2-r_1)$. Substituting the result into the formula for voltage U and reorganizing the formula gives



$$U = N\mu_0 \ell (2\pi)^{-1} \cdot \log_e[r_2/r_1] \cdot \mathrm{d}I/\mathrm{d}t.$$

Now,

$$I = i \sin[2\pi f t]$$

and

$$U = u \cos[2\pi f t],$$

where i and u are the instantaneous values of current and voltage respectively.

$$\therefore\ u \cos[2\pi f t] = N\mu_0 \ell f i \cos[2\pi f t] \cdot \log_e[r_2/r_1].$$

Since it does not matter whether the current and voltage are sinusoidal or not, the term $2\pi ft$ may be ignored, so that

$$u = N\mu_0 \ell f i \cdot \log_e[r_2/r_1].$$

Fig. 3. Note the star connections for the earth lines and the power lines. These are necessary to allow the power meter processing very small signals.

Start construction by laying all the wire bridges: remember that JP_1 and JP_3 must also be laid as such (connection Y in Fig. 2).

Inductor L_2 consists of 40 turns 0.3 mm dia. enamelled copper wire on a G2-3/FT16 core (available from C-I Electronics, see p. 27) to ensure frequency coverage from 20 Hz to 20 kHz. Do not use thicker or thinner wire, because that would change the resistance of the coil, which, in turn, would affect the operation of the integrator.

In view of the sensitivity of M_1, it is essential that the power supply for the power meter is regulated. This may be derived from the amplifier into which the instrument is built or it may be added to the present design with the use of a 7815 and a 7915.

It is impossible to say for all applications what the dimensions of transformer L_1 should be. However, the formula for the relation between u and i in the box on page 10 enables a good-quality secondary to be wound. The fixed data to be used in the formula are $4\pi \cdot 10^{-7}$, i.e., the magnetic permeability of air, u, and f. The latter two must be fixed, because the output of the integrator must not exceed 12 V (that is, approach the supply voltage). This means that at 20 kHz (when L_1 supplies maximum voltage), the peak value of the voltage should not exceed 3.5 V (which is amplified in the integrator by ×3.5). Entering these values and rearranging the formula gives

$$N\ell \cdot \log_e[r_2/r_1] = 139/i.$$

Remember that i is the peak current in amperes to be measured. Thus, if the audio power meter is to be built into a specific amplifier, i is the maximum current the amplifier can provide. Entering that value into the above result leaves only the dimensions and the number of turns to be decided. In principle, the dimensions are dictated by the available space; nevertheless, bear in mind that wrongly chosen values for r_1 and r_2 can have quite an effect on the length and the number of turns. To keep the winding as small as possible, choose as small a size for r_1 as feasible; r_2 should be ×1.5 to ×3 as large as r_1. In view of the logarithm in the formula, a larger or smaller ratio is not sensible. To keep r_1 small, take wire with a thin insulation for the 'primary', for instance, 1.5 mm diameter enamelled copper wire. For the prototype, an empty

PARTS LIST

Resistors:
R1 = 47 Ω
R2, R3 = 1.5 kΩ
R4 = 100 Ω
R5 = 8.2 kΩ
R6, R11 = 22 MΩ
R7, R12 = 4.7 kΩ
R8, R13 = 374 kΩ, 1%
R9, R14 = 1.5 MΩ
R10 = 82 kΩ
R15 = 14.7 kΩ, 1%
R16 = 10 kΩ
R17 = 1 kΩ
R18 = 84.5 kΩ, 1%
R19 = 10 MΩ
R20 = 4.7 Ω
P1 = 100 Ω preset
P2 = 1 kΩ preset
P3, P4 = 100 kΩ preset
P5 = 250 kΩ preset
P6 = 25 kΩ preset
P7 = 2.5 MΩ preset

Capacitors:
C1 = 15 nF
C2 = 560 pF
C3 = 47 pF
C4, C5, C7 = 10 μF*
C6 = 2.2 μF, 100 V*
C8 = 680 nF
C9–C16, C18, C19 = 100 nF
C17 = 47 μF, 40 V
C20, C21 = 100 μF, 25 V
* MKT type

Inductors:
see text

Semiconductors:
D1, D3, D5 = BAT85
D2, D4 = 1N4148
IC1 = LT1028
IC, IC3 = SSM2110
IC4 = AD711
IC5 = AD712

Miscellaneous:
JP1, JP2 = jumper
JP2, JP4, JP5 = 3-way header with jumper
M1 = moving-coil meter, 50 μA (R_i ≈ 3 kΩ)
PCB Type 930018

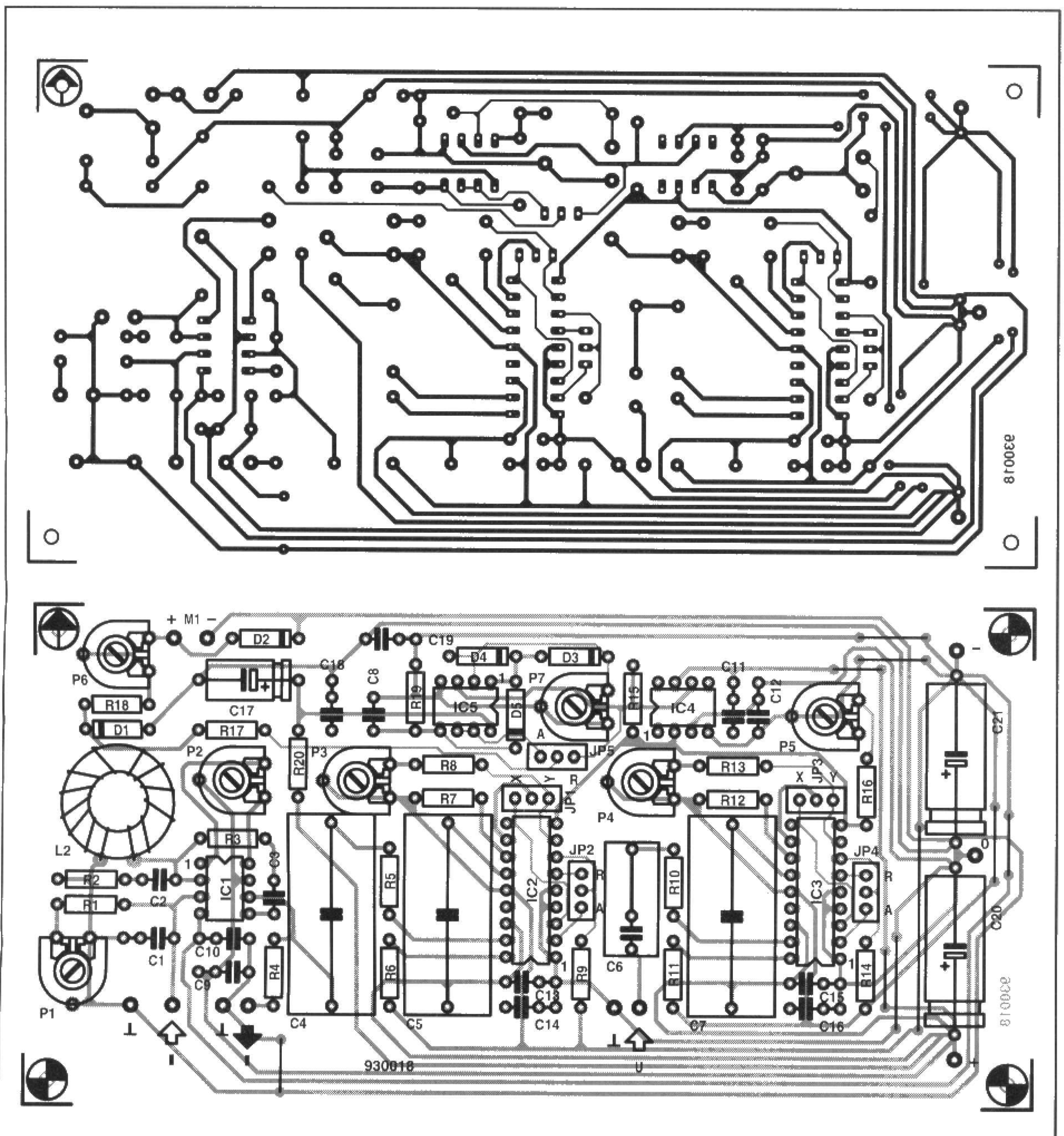


Fig. 3. Printed-circuit board for the audio power meter

solder reel (dimensions: ℓ = 50 mm; r_1 = 7 mm, and r_2 = 25 mm) was used. For a peak current of 28 A, the number of turns required is 77 (round off downwards to prevent difficulties for the integrator). The diameter of the wire is not important, but it is practical to use 0.3 mm diameter, since that is also required for L_2. It should be not too thin to prevent the internal resistance of L_1 becoming too large relative to R_1+P_1.

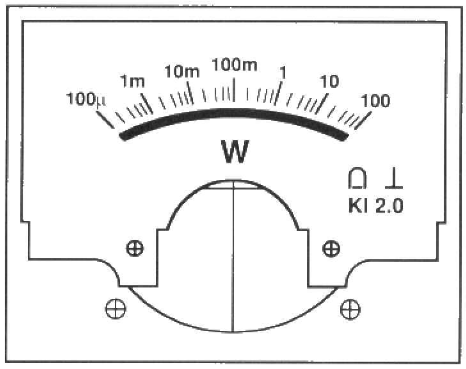
Calibration

Apart from L_1, R_{10} may also be matched to the amplifier, but this is sensible only if the power meter is to be used with that amplifier only. The specified value of the resistor allows a potential of 120 V to be applied, that is, 900 W into 8 Ω. The input range of IC_3 is sufficiently wide to allow low-power amplifiers to be used without having to alter the value of R_{10}. Nevertheless, it is better for the operation of IC_3 to match R_{10} to the amplifier (if this delivers more than 120 V, the resistor MUST be adapted). The correct value of R_{10} can be computed if the value of the peak voltage delivered by the amplifier is known. For convenience, this may be taken as the supply voltage (half the supply voltage if the supply is asymmetrical).

$$R_{10}=U_{peak}/1.5 \quad [\Omega].$$

Round off the result to the next higher standard value.

Before switching on the supply voltage for the first time, set all presets to the centre of their travel and the moving-coil meter to zero (minimum power) with its adjustment screw. Switch on the supply and commence the calibration by nullifying the offset of IC_1 with the aid of P_2. This is done with a mV meter connected to the *I* output that is linked to the integrator. Without any input signal, adjust P_2 until the meter reading is 0 mV or, if that is not possible, very nearly 0 V.



930018 - 15

Fig. 4. Meter scale based on the calculations given in the text.

Use an oscilloscope with one of its channels connected to the *I* output to adjust P_1. Use the other channel to monitor the current through the load. The preset is adjusted at a frequency of 20 kHz such that there is no phase difference between the two signals, or that they are in anti-phase (180°). If an oscilloscope is not available, set P_1 to the centre of its travel. In that eventuality, the power meter may have a slight deviation at frequencies of around 20 kHz. However, since most power is dissipated at low frequencies, that is not terribly serious. If during calibration P_1 needs to be set to maximum resistance, increase the value of R_1; if to minimum (0), reduce the value of R_1.

Presets P_3 and P_4 may be left at the centre of their travel; one of them may be used for fine adjustment of P_5.

The meter range of M_1 is determined by presets P_5 and P_6: because of the log converters, it is largely independent of the values of the input circuits (except L_1, as discussed earlier). First, determine the power that should give full-scale deflection (f.s.d.) of M_1. If the meter is to be used as a stand-alone unit, it is best, from a logarithmic point of view, to use round values, such as 100 W, 300 W, or 1000 W. The other end of the scale is determined by dividing the f.s.d. by 1 000 000; if this results in less than 100 µW, divide by 100 000 to prevent noise being measured.

Once the lowest and the peak values of the meter range have been determined, P_5 and P_6 can be set. Do this with the jumpers in position R. At minimum power adjust P_5 till the meter reads 0 and then, at maximum power, adjust P_6 for f.s.d. It is possible, in case of difficulty with obtaining maximum power, to set P_6 at lower than maximum power.

If minimum power poses a problem, powers of, say, 10 mW and 10 W may be used to adjust P_5 and P_6 respectively. This has, however, the disadvantage that the two presets influence one another. Start by calculating the deflection of the meter as a percentage of f.s.d. at the chosen powers (*P*) with the following equation. Note that this may also be used to calculated the values to be entered on the scale—see Fig. 4.

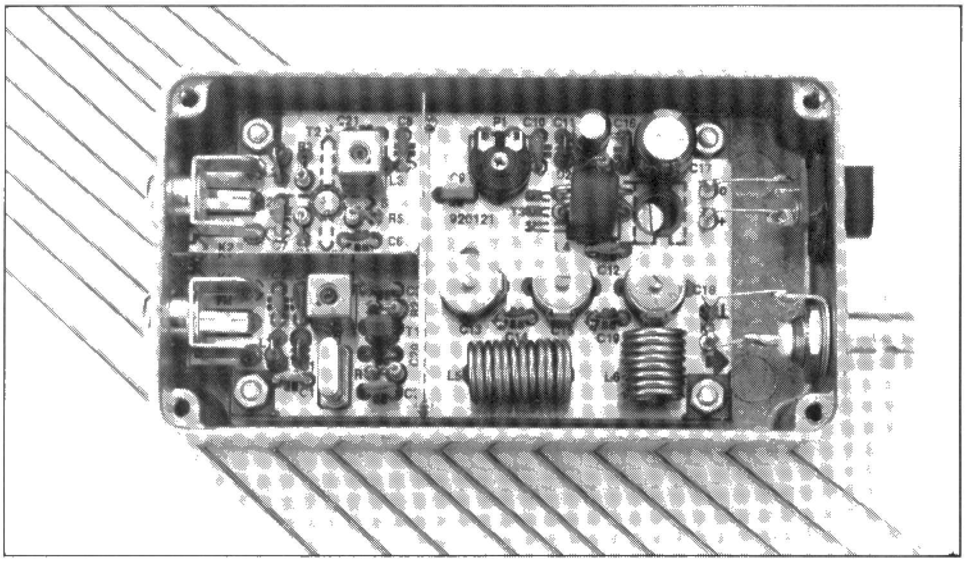
$$\text{deflection}=\log_{10}(P/P_{min})/\log_{10}(P_{max}/P_{min})\cdot 100\%.$$

Then set P_5 to obtain the calculated deflection at low power, and P_6 to obtain the calculated deflection at high power. Repeat setting them a couple of times (in practice, about 3–4) until the presets need not, or hardly, be readjusted.

Finally, calibrate the peak detector, which is done at minimum power and with jumpers JP_2, JP_4 and JP_5 in position A. Set P_7 so that M_1 reads minimum power (0 V at the output of IC_{5b} and pin A of JP_5).

That completes the calibration. Jumpers JP_2, JP_4 and JP_5 can now be set to the desired position (all three in the same position, though). ■

27-MHz AM/FM TRANSMITTER



This article describes a 3-transistor test transmitter for the popular 27-MHz (11-metre) band, which is used for short-range communication ('Citizen's Band') as well as for radio controlled models.

Original design by U. Kunz

THE 27-MHz or Citizen's Band has two major groups of users: radio control (R/C) modellists and users of low-power FM transceivers for local communication. The equipment used by both groups is subject to licensing by the national PTT authorities (Department of Trade and Industry in the UK). The licensing is co-ordinated at an international level by the CEPT (Commission Europeenne de Postes et Telegraphe), while the frequency allocations are provided by the WARC (World Administrative Radio Conference). In most European countries you do not need to pass an examination to obtain a CB license. However all CB transceivers must be type-approved, and may not be modified in any way. In addition, there are strict regulations as regards transmit power, modulation type (narrow-band FM), antenna size and frequency use. Most CB communication is short-range (typically up to 10 km), and concentrated in and round large cities and on motorways, mobile communication being allowed also.

The transmitter described here is intended for testing antennas and aligning receivers. It is quartz controlled for optimum frequency stability, and has an RF output power of about 0.5 watt. Powered by a 12-V supply, it is also suitable for mobile and portable use.

Circuit description

The circuit diagram (Fig. 1) shows a classic three-transistor transmitter design based on FETs (field-effect transistors). The oscillator built around FET T1 derives its frequency stability from a quartz crystal, X1. Here, an inexpensive third-overtone series resonance crystal is used. The oscillator is 'forced' to operate on the third overtone of the quartz crystal by adjusting the L-C parallel tuned circuit in the drain line to 27 MHz. Capacitor C20 is required to ensure sufficient feedback in the oscillator, and also improves its start behaviour. Frequency modulation at a low deviation (NBFM) is achieved with the aid of a variable-capacitance diode ('varicap'), D1. The audio input signal (150 mV_{pp} max.) is applied to connector K1.

The oscillator signal induced in the secondary winding of L1 is fed to the gate-1 terminal of MOSFET T2, a BF982. Gate 2 of T2 is held at about half the supply voltage by R2-R3 to achieve maximum amplification. If AM (amplitude modulation; quite rare these days) is required, the modulation signal may be connected to K2 via a coupling capacitor. The audio voltage will vary the gate-2 voltage of the MOSFET, which results in linear (within limits!) gain control of the MOSFET. The effect is an amplitude-modulated RF output signal. An audio level of 130 mV_{pp} results in a modulation depth of about 70%.

The quiescent current of the power amplifier transistor, T3, is set with the aid of preset P1, which determines the gate bias. Note that the supply voltage of the preset is heavily decoupled to prevent supply and zener diode noise interfering with the RF signal at the gate. The RF power transistor is a HEXFET® Type IRF520 from International Rectifier. As shown, the transistor is cooled by a heatsink. The output filter is a classic pi-type low-pass designed to reduce harmonics and match the output transistor to a load of 50-Ω, which is connected to K3.

Construction

The construction of the transmitter is best started by making the inductors. First, concentrate on the coupled inductors, L1 and L3. Study their orientation on the PCB to make sure that the primary and secondary windings go to the right base pins.

- L1: wound on Neosid 7T1S core. Primary (1-3) = 8 turns; secondary (4-5) = 2 turns. Wire: enamelled copper, 0.2 mm dia. (SWG36).
- L3: wound on Neosid 7T1S core. Primary (1-3) = 10 turns; secondary (4-5) = 2 turns. Wire: enamelled copper, 0.2 mm dia. (SWG36).

Use an ohmmeter to check the continuity of the windings at the base pins. Do not mount the ferrite cup and the screening can as yet (Fig. 2). We continue with the inductors in the power output amplifier.

- L4 consists of 3 turns of 1-mm dia. (SWG20) enamelled copper wire through a 2-hole ferrite balun bead. As indicated by the PCB overlay, this inductor is mounted vertically.
- L5 consists of 12 turns of 1-mm dia (SWG20) enamelled copper wire.

Please note that the transmitter described here is not licensable in the UK.

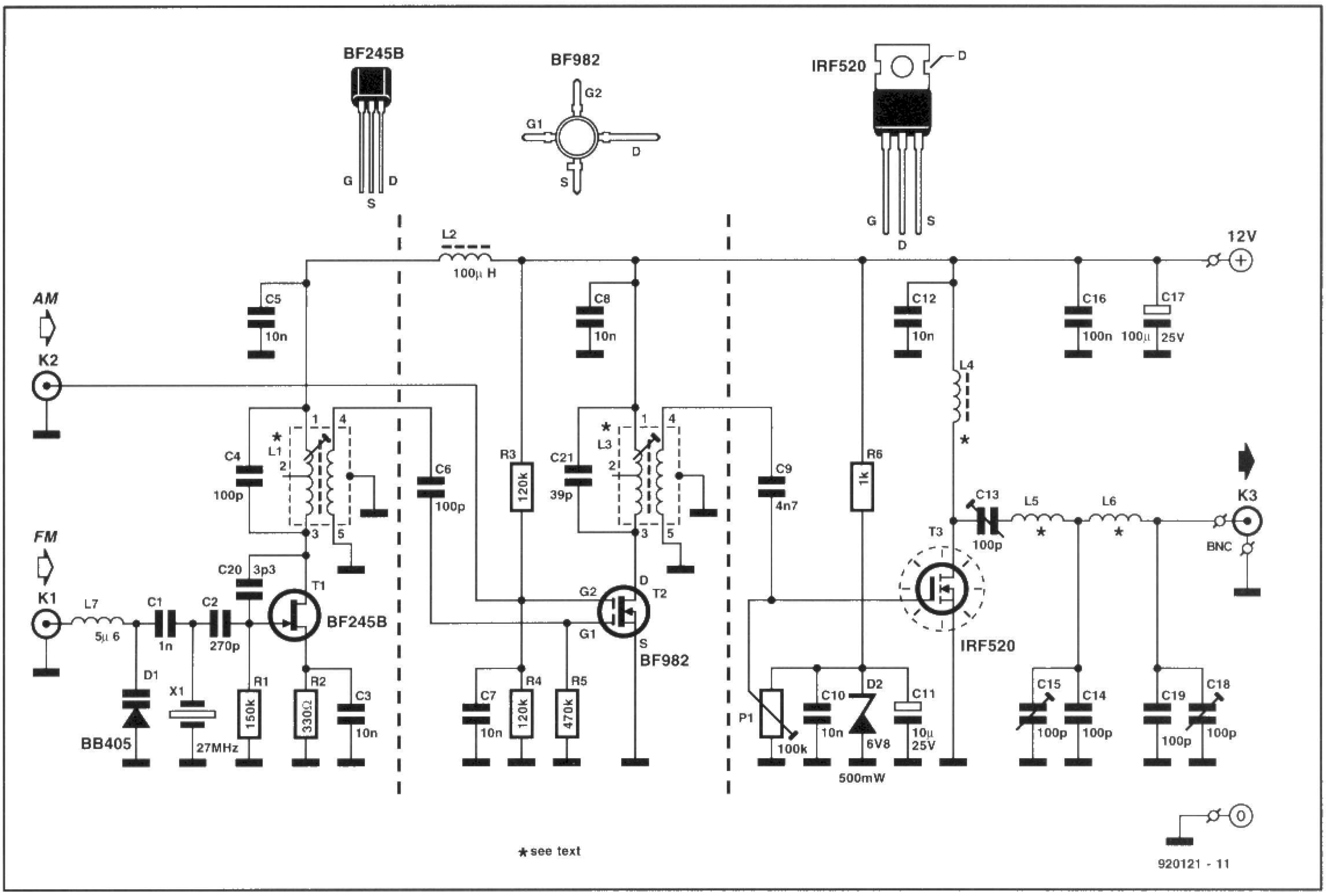


Fig. 1. Circuit diagram of the 27-MHz AM/FM test transmitter.

Closewound, internal diameter 8 mm; no core.

L6 consists of 8 turns of 1-mm dia. (SWG20) enamelled copper wire. Closewound, internal diameter 8 mm; no core.

The PCB design is given in Fig. 3. It should be noted that the board is double sided, but not through-plated. This means that component wires must be soldered at both sides of the board where possible. Also, all component wires must be kept as short as possible.

Start by mounting inductors L1 and L3. Do not mount the screening cans as yet. As indicated by their dashed outlines on the PCB overlay, transistors T2 and T3 are fitted at the underside of the board. This allows T3 to be secured to the bottom of the metal enclosure in which the PCB is fitted later. Do not forget to use an insulating washer, since the metal tab of the IRF520 is connected to the drain. The type indication of T2 is legible from the top side of the board.

The remainder of the construction is all plain sailing, and should not cause problems if you have some experience in building radio projects. The audio input sockets are PCB-mount types. The oscillator, buffer and power amplifier are screened from one another by pieces of 15-mm high tin plate sheet fitted vertically on the dashed lines on the PCB overlay.

As shown by the introductory photograph of our prototype, the board is fitted in a diecast case. Although a BNC socket was used on the prototype, an SO-239 type is equally suitable for the RF output. The DC power supply input is formed by a two-way adaptor socket as used on portable radios and cassette recorders.

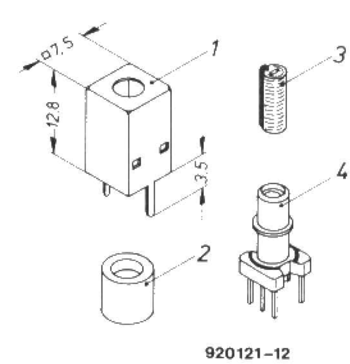


Fig. 2. Neosid 7T1S inductor assembly. (1) screening can, (2) ferrite cup, (3) core, (4) former and base.

Adjustment

You will need the following equipment to adjust the transmitter: a frequency meter or a grid-dip meter, a dummy load or an in-line SWR/power meter, an insulated trimming tool, and a regulated 12-V power supply. Mount a small TO-220 style heat-sink to the tab of T3.

First, turn the wiper of P1 to ground, and set the three trimmers roughly to mid-travel. Insert the cores in L1 and L3. There is no need to apply a modulation signal as yet to either input. Apply power, and couple the frequency meter or GDO inductively to L1. Adjust the core until the oscillator operates at the quartz crystal frequency. Switch off and on again to check the start behaviour. Move on to L3, and adjust the core for resonance at 27 MHz. This is readily measured by moving the pick-up device further away from the inductor. If you can not seem to find a clear maximum ('peak') in this way, do not worry, as this is only a provisional adjustment. Next, keep a close eye on the current consumption of the transmitter. Carefully

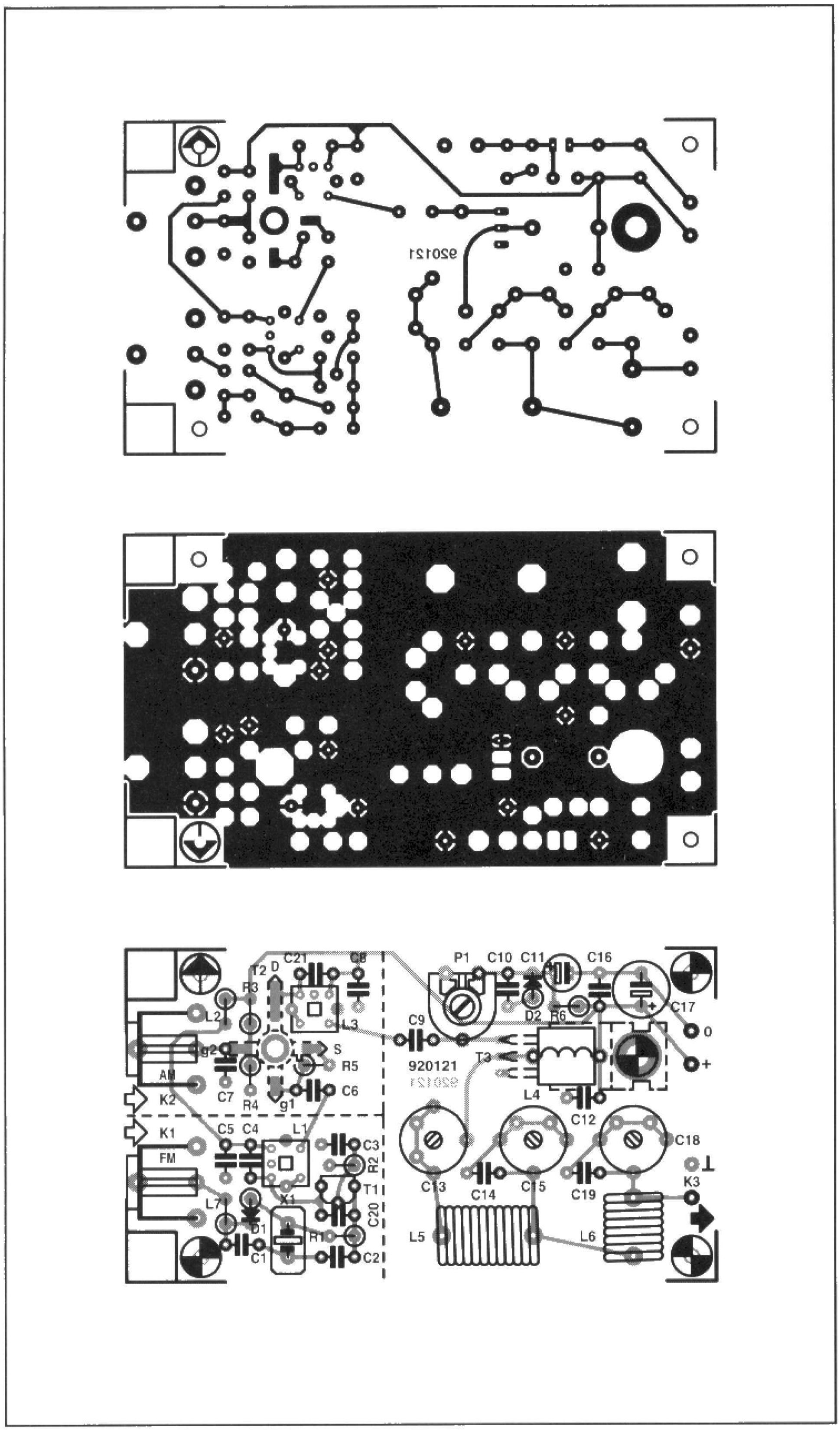


Fig. 3. Artwork for the double-sided, not-through plated PCB. This PCB is unfortunately not available ready-made through the Readers Services.

COMPONENTS LIST

Resistors:

1	150kΩ	R1
1	330Ω	R2
2	120kΩ	R3;R4
1	470kΩ	R5
1	1kΩ	R6
1	100kΩ preset H	P1

Capacitors:

all ceramic capacitors 5mm raster

1	1nF ceramic	C1
1	270pF ceramic	C2
6	10nF ceramic	C3;C5;C7; C8;C10;C12
4	100pF ceramic	C4;C6;C14;C19
1	4nF7 ceramic	C9
1	10µF 25V radial	C11
3	100pF foil trimmer	C13;C15;C18
1	100nF ceramic	C16
1	100µF 25V radial	C17
1	3pF3 ceramic	C20
1	39pF ceramic	C21

Inductors:

2	7T1S assembly (Neosid) for winding details see text	L1;L3
1	100µH choke	L2
1	2-hole ferrite balun bead; size approx. 15×8×8 mm*	L4
1	5µH6 choke	L5

0.2mm dia. enamelled copper wire
1mm dia. enamelled copper wire

Semiconductors:

1	BB405	D1
1	6V8 0.4W zener diode	D2
1	BF245B	T1
1	BF982	T2
1	IRF520**	T3

Miscellaneous:

2	PCB-mount audio socket	K1;K2
1	27MHz quartz crystal	X1
1	BNC or SO-239 socket	K3
1	Diecast case 112×61×28mm, e.g. Hammond 1590B or Cricklewood DBX30	

* Bonex Ltd. (0753) 549502. Part no. 26-43006301.
** Cricklewood Electronics (081) 452 0161.

adjust P1 until the current drain is about 100 mA, and monitor the output power. Peak the three trimmers for highest output power. The trimmer adjustments will interact slightly, so you may have to spend some time before the optimum settings are found. Next, fine-tune L3 for highest output power.

Finally, fit the ferrite cups and the screening cans on L1 and L3. After removing the provisional heat sink on T3, the completed board is fitted into the enclosure. This is done with the aid of PCB spacers and screws, for which there are four PCB corner holes. T3 is secured on to the bottom of the case, using a mica washer. The screw is accessible through the hole in the PCB. Use an ohmmeter to check that the tab of the transistor is isolated from the diecast case.

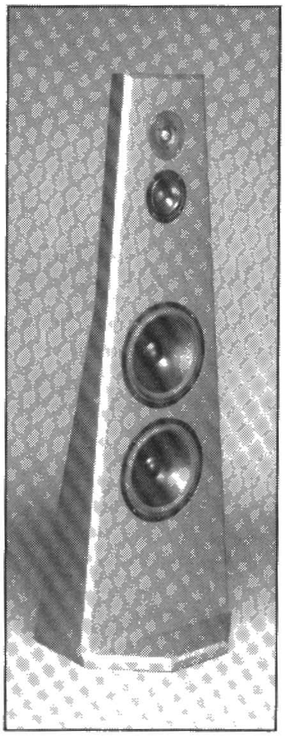
Finally, make sure that preset P1 is set to minimum PA current drain (wiper fully to ground) before applying an AM modulation signal. Carefully adjust P1 for an output power of about 0.5 W PEP (peak envelope power) into a 50-Ω load. ■

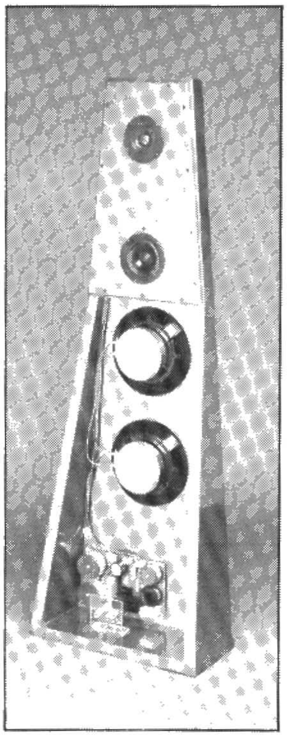
OPEN-BAFFLE LOUDSPEAKER

Design by H. Baggen

Some Technical Data

Type of box	Open with dipole behaviour
Drive units	Two 210 mm woofers
	Two 100 mm squawkers
	Two 19 mm tweeters
Cross-over frequencies	400 Hz and 5 kHz
Power handling	Up to 150 W (music)
Efficiency	86 dB (relative to 1 at 2.83 V input)
Nominal impedance	4 Ω
Dimensions	1175×400×300 mm (H×W×D)
Estimated cost per box	£200–£225, excl. wood





At present, Vifa-Speak, the Danish manufacturers of the drive units, have no representation in the UK or the USA. Their only outlets for the drive units are:

Audio Components BV
Postbus 554
5340 AN Oss
The Netherlands
Phone +31 4120 26610
Fax +31 4120 33017

or

Audio Design GmbH & Co KG
Wuppertalerstr. 12
D-4300 Essen
Germany
Phone +49 201 471 023
Fax +49 201 473 730

Introduction

A design is presented that offers an alternative to the usual loudspeaker enclosure. Its radiation pattern is that of an electrostatic design; it has no enclosure for the woofers and yet it uses standard dynamic drive units. Its reproduction gives a very 'spacious' impression.

Design considerations

Electrical energy is usually converted into acoustic energy by a dynamic drive unit. There are other types, such as the electrostatic and ribbon units, but these are generally more expensive and often also more vulnerable or more difficult to make than the good old cone type that has been around for some 70 years. Worldwide, tens of millions of loudspeaker drive units are produced annually. Only a small portion of these is intended for hi-fi apparatus; the remainder is for use in telephones, car radios, portable radios, and so on.

Only the cone loudspeaker is normally considered for high-quality sound reproduction, because only this type can set a large volume of air into motion (which is essential for the physiological hearing characteristic). When a diaphragm is to reproduce low frequencies, it is essential that the front and rear of its cone cannot 'see' each other (otherwise, an acoustic short-circuit would arise). For this reason, a closed box or bass reflex enclosure is normally used for the reproduction of low frequencies. Such an enclosure has the disadvantage, however, that it tends to move with the cone (except perhaps if it is set in concrete). This aspect will be reverted to shortly.

Another aspect of sound reproduction is the radiation pattern of the loudspeaker system.. Some experts feel that the design should aim to be a point source, that is, all frequencies should be radiated over an angle of 360°. In practice, the radiation pattern of mid-frequency and tweeter units is limited to 180° (or thereabouts); only the woofers will apprach 360°. There are solutions to this, for instance, mounting drive units at the rear of the enclosure also. Another is the use of electrostatic drivers, since these radiate forward and backward. As the reproduced sound at the front is in anti-phase with that at the rear, these units behave in a different way from an omni-directional radiator. Such units are therefore called dipole radiators, although the radiation pattern is octagonal. The sound from such a unit can, however, be very pleasant to listen to, because the sound waves from the rear reach the listener via a number of reflections, which add to the stereoscopic impression.

The present design, which, although not new in the audio world, has never to our knowledge been the subject of a DIY project, aims to combine these various aspects. In other words, one that performs as a dipole radiator but uses standard dynamic drive units. Low frequencies are reproduced by two woofers mounted on a small baffle, while mid-range and high frequencies are reproduced by two squawkers and two tweeters, one of each at the front and one of each at the rear.

Practical considerations

When a loudspeaker drive unit is mounted at the centre of a board, its frequency characteristic below the lower cut-off frequency (which depends on the dimensions of the board) will drop at the rate of 6 dB per octave. Below the resonance frequency of the drive unit that will increase to 18 dB per octave, but this is immaterial in the case of drive units with a low resonance frequency. This response is much better than that of a closed box (12 dB per octave) or that of a bass reflex enclosure (12–18 dB per octave). The disadvantage is, of course, that the lower cut-off frequency is high (half wavelength = diameter of board). At this frequency, the front and rear of the cone begin to counter one another so that the resulting performance deteriorates. It is as if the air pushed forward at the front is absorbed by the air sucked in by the cone at the rear. A cut-off frequency of 60 Hz requires a board of almost 3×3 m (10×10 ft). Moreover, a smooth characteristic requires the drive unit to be mounted asymmetrically, so that the 'short circuits' are spread over a wide frequency range. Such a large board is, of course, out of the question for domestic use, whence the designs for similar performance that take up much less space. Nevertheless, the open baffle design remains of interest, because it precludes the many (un-

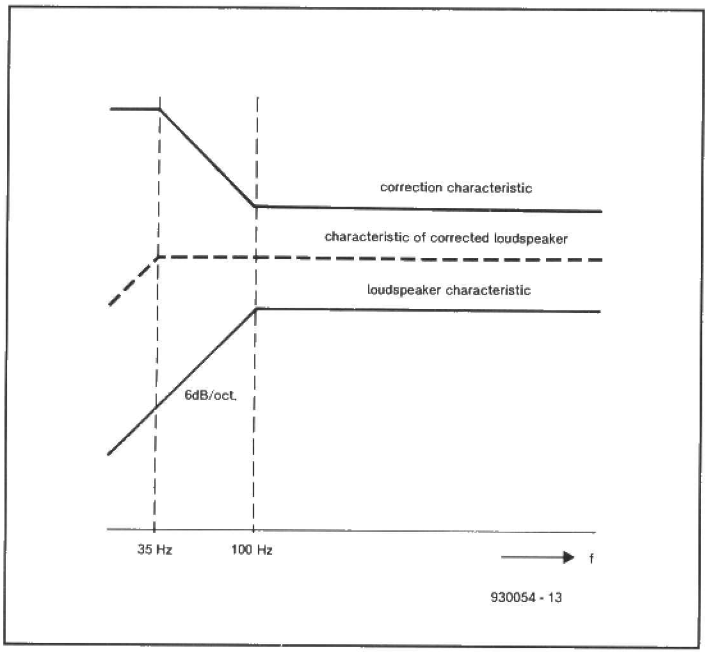


Fig. 1. Calculated correction and loudspeaker characteristics and the resulting characteristic.

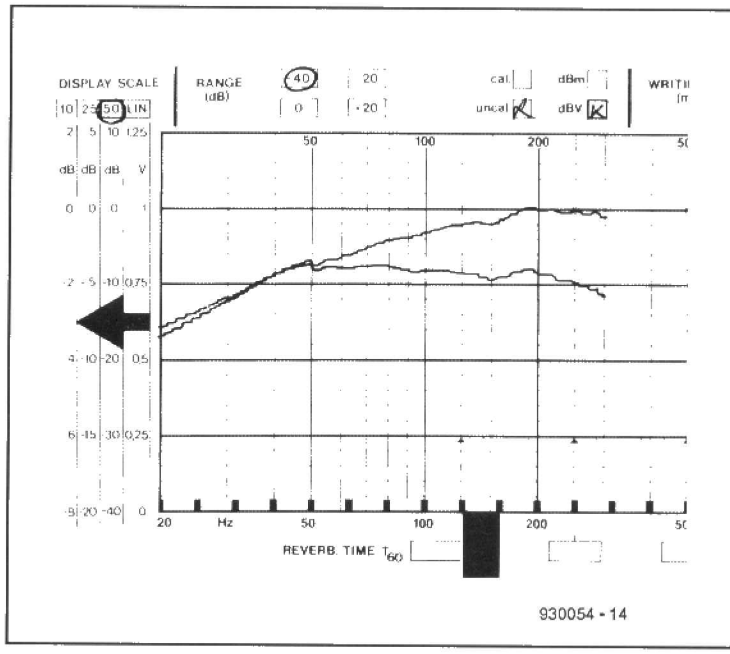


Fig. 2. Measured characteristics of uncorrected loudspeaker (upper curve) and corrected loudspeaker (lower curve).

desirable) effects an enclosure has on the reproduction of sound (standing waves; vibrating in unison, and so on). Vibration of the enclosure is a particular problem when the box is made of wood.

For domestic use, a board of modest dimensions placed on the floor of a room could be used to increase its size artificially and thus lower the cut-off frequency. Also, electrical compensation may be used to (partly) compensate the acoustic deterioration. This will reduce the efficiency and power handling to an extent, but this can be kept within reasonable limits by the use of a large cone and restricting the correction. The present design uses a high, narrow board on which two 210 mm woofers are mounted and is intended for standing on the floor. The (measured) low cut-off frequency (–3 dB) lies at around 100 Hz.

Since an extra amplifier was deemed unnecessary, the correction network is a passive *LC* type connected at the input of the woofers—see Fig. 3. The (calculated) characteristic of the woofer mounted on a board, that of the correction network, and that of the corrected loudspeaker are shown in Fig. 1. To retain the efficiency and power handling within acceptable limits, the correction was limited to just over one octave. The efficiency is reduced by 8 dB. The use of two woofers does not really increase the efficiency (the total impedance is lower but the dissipation is higher), and the power handling remains virtually the same as that of one 210 mm woofer in a closed box. The measured characteristics are shown in Fig. 2. It is seen that the –3 dB cut-off frequency is lowered to around 35 Hz, which is a good value for hi-fi applications. Note that the corrected curve includes the effects of the low-pass filter, so that it starts to fall off again above 200 Hz. The resulting design is a narrow baffle that gives good bass performance even at low frequencies. Compared with some 'normal' boxes, it may appear as if the present design does not reproduce the low frequencies so well, but that is because box designers often do not allow for the effect of the actual room, which results in a low frequency peak once the loudspeaker is used in the room.

The characteristics of the woofers measured at the front and rear are virtually identical at low frequencies. This is not the case with squawkers (mid-range units) and tweeters, which means that these have to be duplicated at the rear. Furthermore, squawkers have a severaly curved frequency characteristic and poor radiation performance. Because of that, it was found necessary to place these units in a small enclosure.

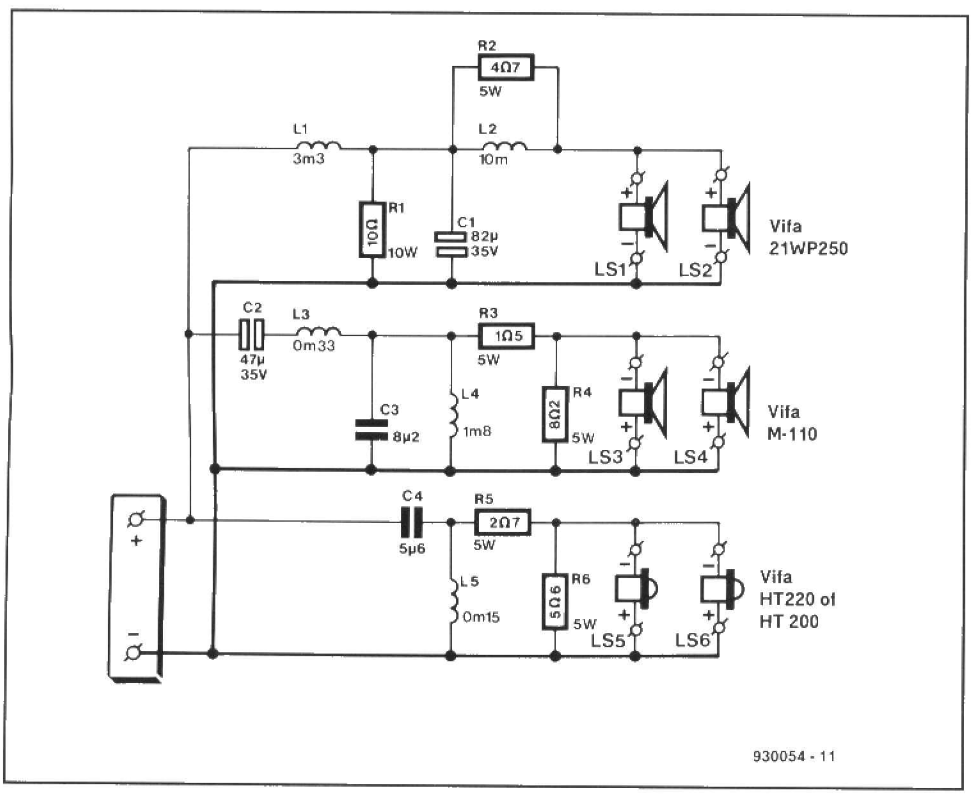


Fig. 3. Circuit of the cross-over filter for the open-baffle loudspeaker.

Choice of drive units

To obtain good radiation performance, the diameter of the drive units must be

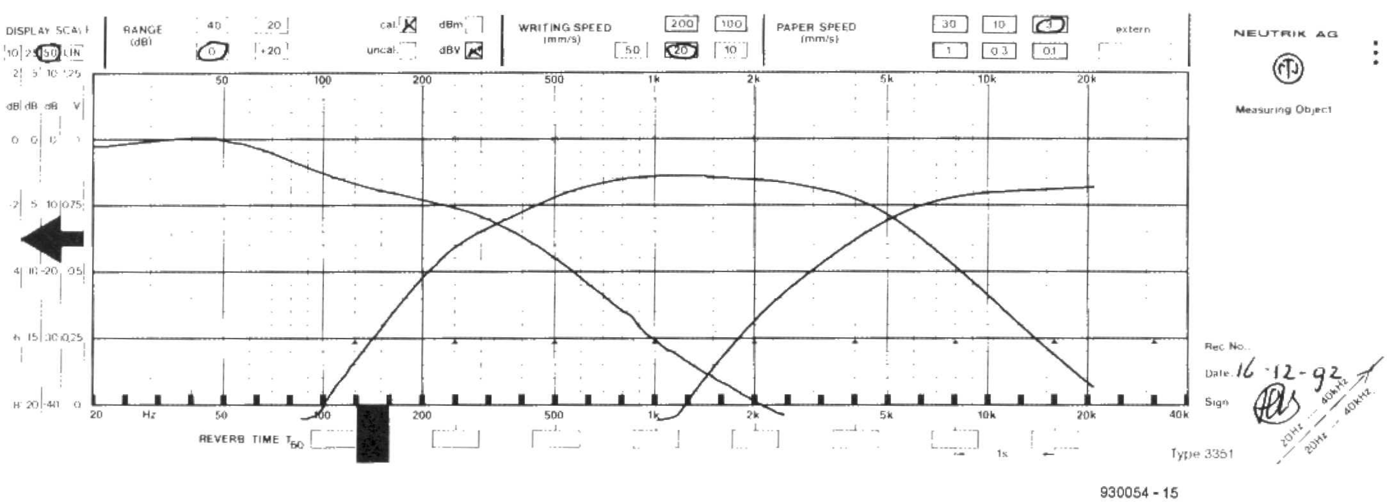


Fig. 4. Electrical output signals of the three cross-over network sections.

small compared with the wavelengths of the reproduced sounds. This means that a three-way system is required. Since the choice of different types of drive unit in a system often gives rise to availability problems, it was decided to select all three types from the range of one manufacturer. Because of their good price-to-quality ratio, units from the Danish producer Vifa were chosen.

The Type 21WP250 has a cone of man-made fibre. Owing to its large magnet, the unit exhibits strong electrical damping, so that, in spite of the absence of an enclosure, good pulse behaviour obtains (in an enclosure, the air acts as damper).

The Type M-110 mid-range unit has a cone of coated paper. A double magnet functions as the driver, but in spite of that the unit is small enough to be placed in a box of just one litre.

The tweeter may be a Type HT220, which has an aluminium dome or, if a soft dome is preferred, the Type HT200. The types are fully interchangeable: the cross-over filter remains the same.

The cross-over network

The circuit of the cross-over network is shown in Fig. 3; its measured output in Fig. 4. Inductor L_2 and R_2 provie the low-frequency correction shown in Fig. 1. Filtering proper is provided by L_1-C_1. This section provides a second-order slope above about 400 Hz (it appears to be somewhat lower in Fig. 4, but that is because the curves there pertain to the electrical output only: the efficiencies of the drive units are not included). Resistor R_1 ensures a reasonably constant resistance at the output of the network, in spite of the effect of L_2-R_2 and the frequency-dependent impedance of the woofers.

The section for the squawkers consists of L_4-C_2 for the roll-off at 400 Hz and of L_3-C_3 for that at 5 kHz. The slopes are about 12 dB per octave. Together with the natural roll-off of the tweeters, this results in a steeper slope, which is essential to ensure that the mid-range units do not handle too much power. Attenuator R_3-R_4 between the section and the drive units provides level matching at around 3.5 dB.

The tweeter section (second-order) consists of L_5-C_4. Attenuator R_5-R_6 provides level matching at around 5.5 dB to give a final, flat acoustic frequency response.

The cross-over network is best built on a piece of prototyping (Vero) board—see Fig. 5 for a suitable component layout. The cored inductors are fairly heavy and must, therefore, be secured properly, preferably with non-metallic nuts, bolts and washers. Standard 2-way terminal blocks are used for the various cable con nections.

Inductors L_1, L_2, and L_4 are bobbin types with an HQ core. This material does not cause distortion at low and high levels alike and is relatively inexpensive. Since L_1 and L_2 handle fairly large currents, do not use bobbin inductors with unknown or inferior core material.

Although C_2 is specified as a bipolar electrolytic, it may also be an MKT type.

Construction

Virtually all the panels shown in Fig. 6 are made from 25 mm (1 in) medium densitiy chipboard. The major component is panel A, a 1150 mm high board on to which two woofers, a squawker and a tweeter are mounted. Note that all drive units must be fitted into sunk holes, which considerably enhances their radiation performance. This is, however, only necessary at the front, since the radiation from the rear is not that impor-

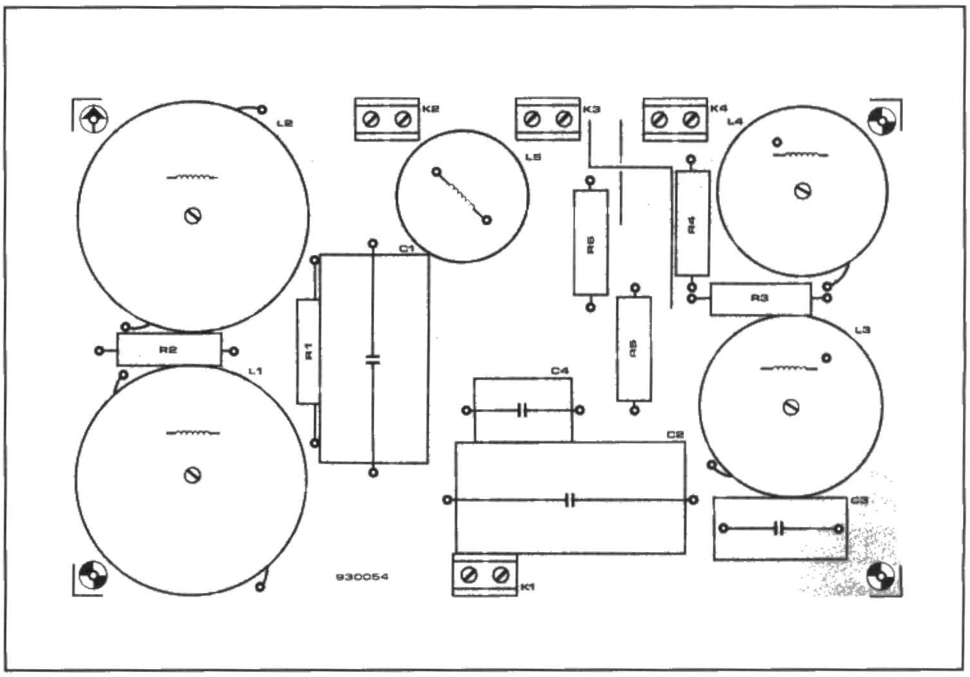


Fig. 5. Suggested component layout for the cross-over network (scale 1:2).

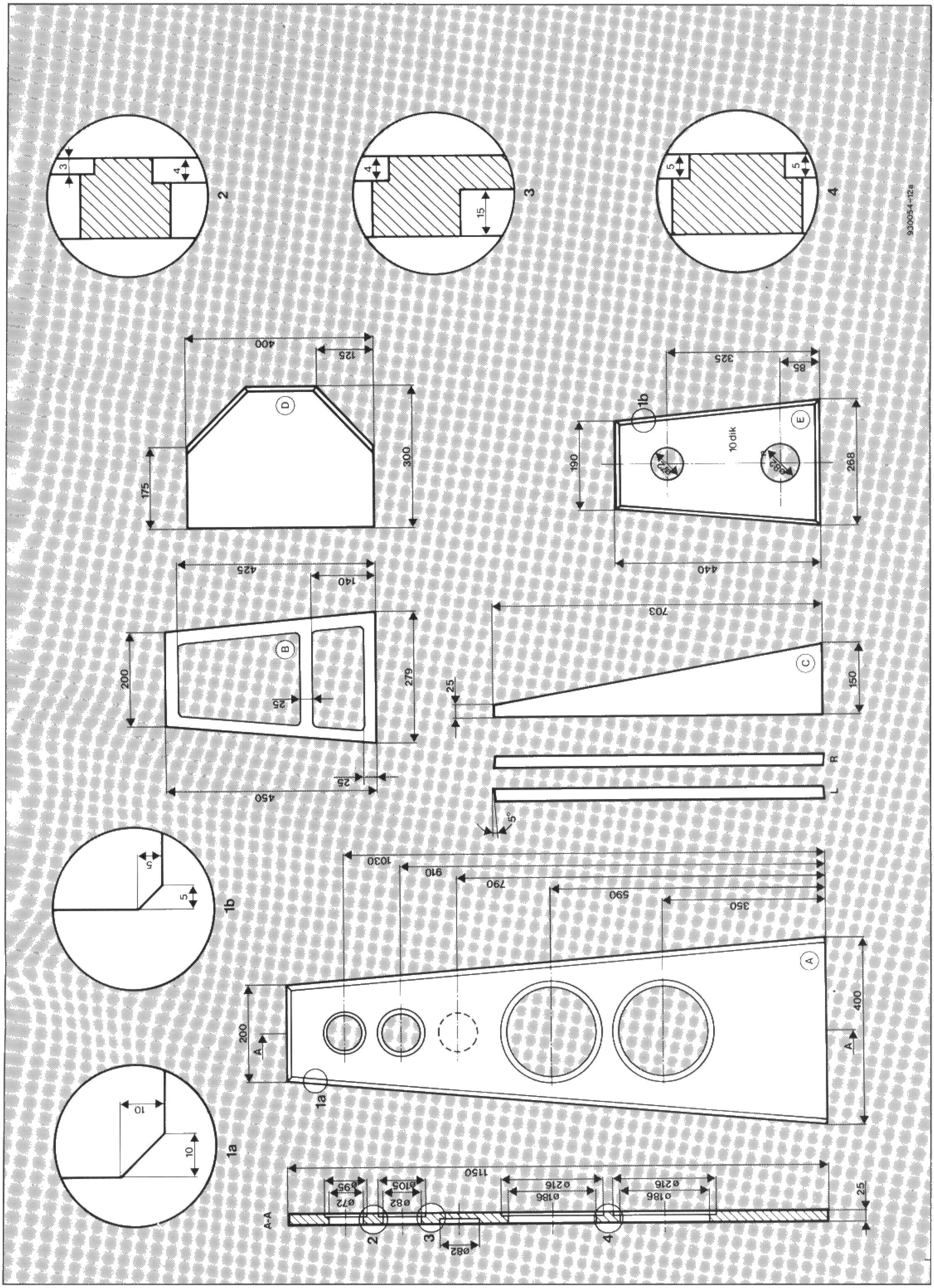


Fig. 6a. Panels and constructional details of the open baffle.

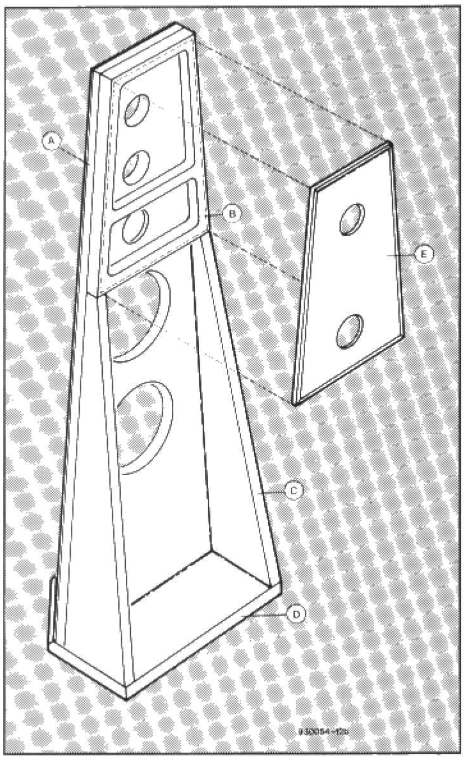


Fig. 6b. The nearly finished baffle.

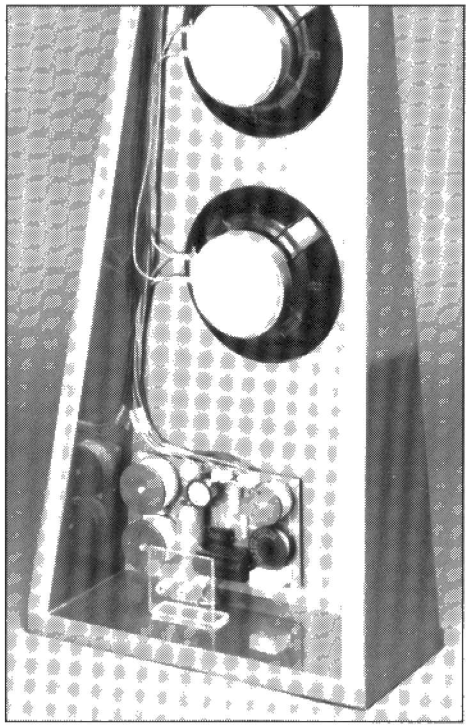


Fig. 7. Position of cables, cross-over network and L-shaped bracket.

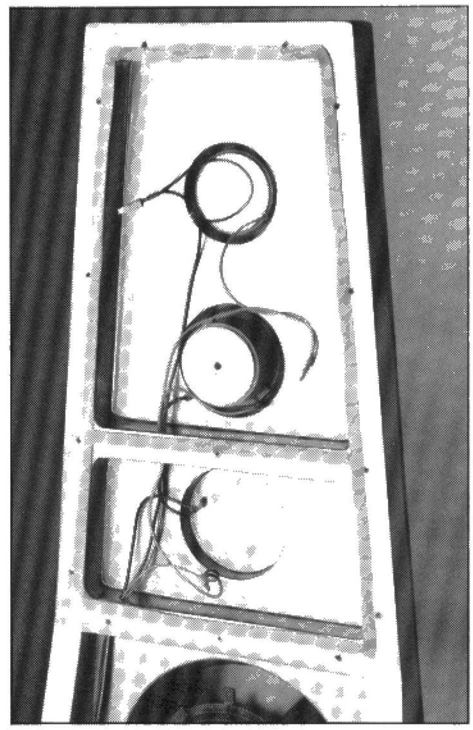


Fig. 8. Rear view of the nearly finished baffle; note the milled hole for the squawker.

tant. One of the trickiest operations is the milling of a 15 mm ($^{19}/_{32}$ in) deep hole at the rear of the board into which later the magnet for the rear squawker is to rest—see Fig. 8.

The large board is fixed to the base (D), after which a support (C) is fitted to each of its sides. It is advisable to reinforce the structure with some screws through the base into panels A and C.

When the holes as shown have been cut into panel B, drill two 8 mm holes into the lower horizontal struts (for the cables–if these are thinner, the diameter of the holes may be smaller).Next, glue the panel to the top rear of panel A—see Fig. 6b. This creates the box for the squawkers and tweeters.

A sideview of the construction so far looks like a single, 50 mm (2 in) thick panel that broadens towards the bottom. Finish this as required in laquer or veneer. Do not forget panel E

When the laquer or veneer is dry, run the cables for the loudspeakers and fit the drive units at the front—do not forget the cables for the rear squawker and tweeter. Connect the cables to the units. Mark the cable ends, so that it is clear later on, when the top box is closed, which cable goes to which unit. The holes through which the cables run must be closed airtight with, e.g., a glue gun.

Next, fasten panel E with chipboard screws to the top rear as shown in Fig. 6b. The screwheads must be sunk. Cover the gaps between the panels with suitable tape. Then mount the rear squawker and tweeter. Make sure that the connections to these units are mirror images of those at the front; that is, connect the + line of the front tweeter to the – line of the rear tweeter, and similarly with the mid-range units. The electrical polarity relative to the cross-over network is determined by the front loudspeakers.

Then, mount the cross-over network below the woofers as shown in the photograph in Fig. 7.

Finally, make an L-shaped bracket as shown in Fig. 7, screw this to the base panel as shown and fit the sockets to this. Connect the sockets to the cross-over network as appropriate.

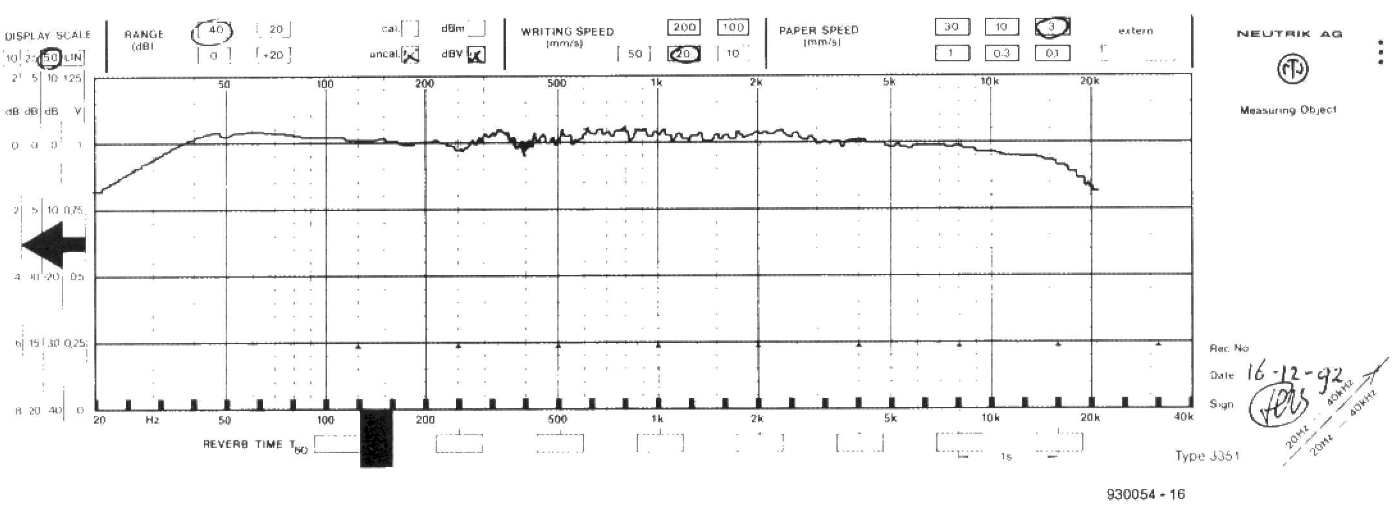


Fig. 9. Frequency response measured at the front of the open baffle. That at the rear is virtually identical.

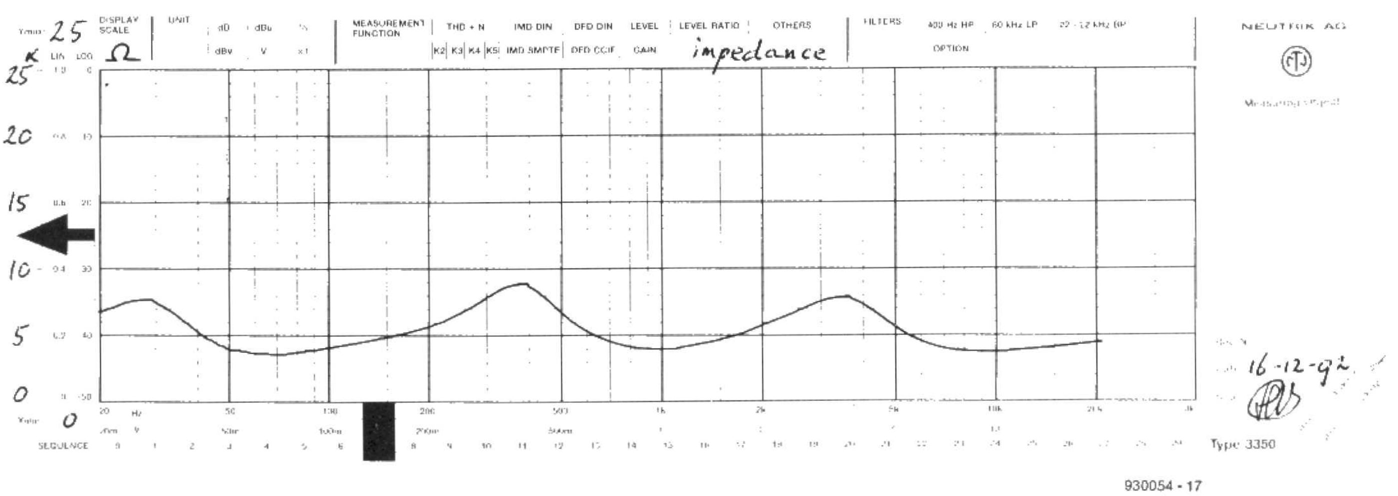


Fig. 10. Impedance characteristic of the open baffle loudspeaker system.

Test results

The frequency response curves measured at the front of the prototypes (above 250 Hz at 1 m distance, below that frequency at 10 cm from the woofers) are shown in Fig. 9.

The impedance curve of the system is shown in Fig. 10. Since the minimum impedance is 3.5 Ω at 70 Hz, most amplifiers will work very well with the loudspeaker, particularly since the impedance variations over the entire frequency range are small (3.5–9 Ω), which limits any phase shifts.

Experiment with the location of the open-baffle loudspeakers in the room. There must be sufficient space at their back: this makes these speakers unsuitable for small rooms. Also, they must be away from side walls at least 1 m. The location of the speakers relative to the walls determines the nature of the reflections from the rear speakers as well as the power of the low-frequency sounds. The bass is enhanced when the speakers are placed closer to the wall behind them. Reflections at mid and high frequencies depend very much on the nature of the walls and on the layout of the room. If these frequencies are reproduced somewhat too weakly or too strongly, increase or reduce, respectively, the values of R_3 (mid frequencies) or R_5 (high frequencies) by an E value. This will have no noticeable effect on the cross-over network.

As far as sound is concerned, one of the most notable properties of the open-baffle speakers is the spacious character. This does not always means an improvement of recorded sound. The impression is that 'live' recordings are enhanced, whereas 'doctored' ones sound worse.

Another aspect is the broad auditorium. Normally, only someone sitting exactly at the centre of the loudspeakers will hear perfect stereo reproduction. With the open-baffle units, you can move to the left or right without losing the true stereo effect.

Finally, the bass response is very faithful. Any spatial resonances could be eliminated by slightly altering the angle of the units. The impression that the bass response is not as good as that of closed box loudspeakers has already been commented on. ■

PARTS LIST
(per open baffle)

Resistors:
R1 = 10 Ω, 10 W
R2 = 4.7 Ω, 5 W
R3 = 1.5 Ω, 5 W
R4 = 8.2 Ω, 5 W
R5 = 2.7 Ω, 5 W
R6 = 5.6 Ω, 5 W

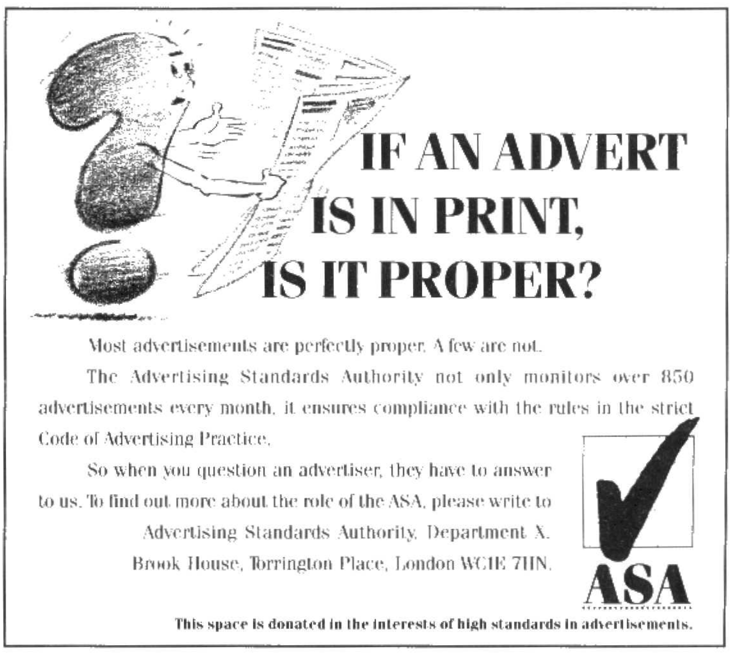
Capacitors:
L1 = 3.3 mH*
L2 = 10 mH*
L3 = 330 μH+, 1 mm wire
L4 = 1.8 mH†
L5 = 150 μH+, 0.71 mm wire

*56 mm high-Q bobbin
+air-cored
†40 mm high-Q bobbin

Drive units:
LS1, LS2 = Vifa 21WP250
LS3, LS4 = Vifa M-110
LS5, LS6 = Vifa HT220 or HT200

Miscellaneous:
Vero board, 200×120 mm ($7\frac{7}{8}$×$4\frac{11}{16}$ in)
4 two-way terminal blocks
Acoustic wadding, about 400×200 mm (16×8 in)
Chipboard, medium density, 25 mm thick:
1 panel (A)1150×400 mm ($45\frac{9}{32}$×$15\frac{3}{4}$ in)
1 panel (B) 450×279 mm ($17\frac{23}{32}$×10 in)
2 panels (C) (703×150 mm) ($27\frac{11}{16}$×$5\frac{29}{32}$ in)
1 panel (D) 400×300 mm ($15\frac{3}{4}$×$11\frac{13}{16}$ in)

Chipboard, medium density, 10 mm thick:
1 panel (E) 440×268 mm ($17\frac{5}{16}$×$10\frac{9}{16}$ in)



FIGURING IT OUT

PART 4 – ANALYSING NETWORKS

By Owen Bishop

This series is intended to help you with the quantitative aspects of electronic design: predicting currents, voltage, waveforms, and other aspects of the behaviour of circuits. Our aim is to provide more than just a collection of rule-of-thumb formulas. We will explain the underlying electronic theory and, whenever appropriate, render some insights into the mathematics involved.

Many network problems can be solved by using the methods described in Parts 1 and 3, particularly by superposition and by deriving Thevenin equivalents. But some networks do not yield to these techniques. Where these fail, **mesh current analysis** may succeed. One point to remember: mesh current analysis applies only to networks with **voltage** sources, not **current** sources.

The idea behind mesh current analysis is very simple. Unfortunately, as the number of meshes increases, the calculations soon become too complex and tedious to handle, and the likelihood of arithmetical mistakes increases out of all proportion. But, because the basic idea is extremely simple, the computations lend themselves to a computer algorithm, whereby arithmetical errors are eliminated. We conclude this part with such an algorithm, transcribed into BASIC, which should run on almost any microcomputer.

The principle of mesh curent analysis is illustrated by reference to the 2-mesh circuit in Fig. 33. Normally, this circuit would be analysed most easily by superposition. To begin the analysis, we specify two **mesh currents**, I_1 and I_2. These are hypothetical currents, flowing in a clockwise direction around each mesh. We could have them flowing anticlockwise, but this would simply alter the signs, making no other difference to the calculation.

Taking mesh 1 to begin with, and ignoring mesh 2, we note that the voltage source is driving the hypothetical current I_1 in the direction indicated by the arrow. Passing around the mesh in the direction of the arrow, we find a voltage **drop**, or negative pd, across each resistor and a voltage **rise**, or positive pd, across the voltage source. Applying KVL, the sum of the voltage drops across the resistors equals the voltage rise across the voltage source.

The total resistance of mesh 1 is 11 Ω, so the voltage drop caused by I_1 is $11I_1$. Looking more closely at the circuit, we see that mesh 1 shares branch CE with mesh 2. There is a voltage drop across this branch caused by the 2 Ω resistor through which current I_2 is flowing. This drop is $2I_2$ but, since I_2 flows through the resistor in the **opposite** direction to I_1, this is a **negative** drop, that is, a voltage rise. Summing the voltages according to KVL:

$$11I_1-2I_2=5. \qquad \text{[Eq. 17]}$$

The same line of argument for mesh 2 produces the equation:

$$-2I_1+5I_2=-4. \qquad \text{[Eq. 18]}$$

The negative signs in this equation reflect the fact that I_1 flows through the 2 Ω resistor in the opposite direction to I_2 and that the polarity of the source tends to drive the current in the opposite direction to I_2.

We now have two unknown variables, I_1 and I_2, and two equations. Since these equations represent the voltages found in the circuit when **both** sources are active, the equations are both true **at the same time**. They are **simultaneous equations**. We have to solve these to find values of I_1 and I_2 that satisfy both equations.

There are several methods of solving simultaneous equations, some of which rely on a certain flair for navigating one's way through the calculations. The benefits of personal flair, if any, are to avoid the awkward fractions that often arise, and to keep the numbers small. All of these solution methods are reducible to relatively simple algorithms. Consequently, a scientific formula calculator often has such a routine as one of its built-in programs. Using such a calculator on equations 17 and 18, we find that the currents through the resistors are:

$$I_1=1/3$$

and

$$I_2=-2/3.$$

Interpreting these results as the currents through individual resistors:

5 Ω and 4 Ω carry $I_1=1/3$ A;

3 Ω carries $I_2=-2/3$ A (opposite direction to I_1);

2 Ω carries $I_1-I_2=1/3-(-2/3)=1$ A (same direction as I_1.

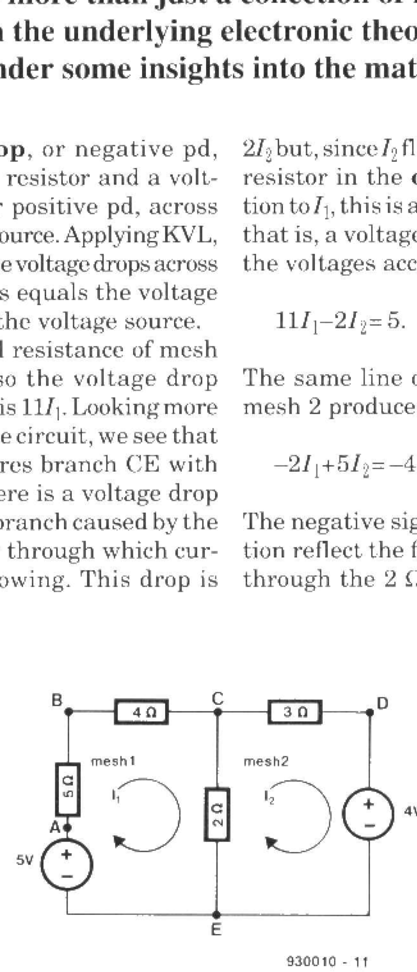


Fig. 33. A 2-mesh circuit.

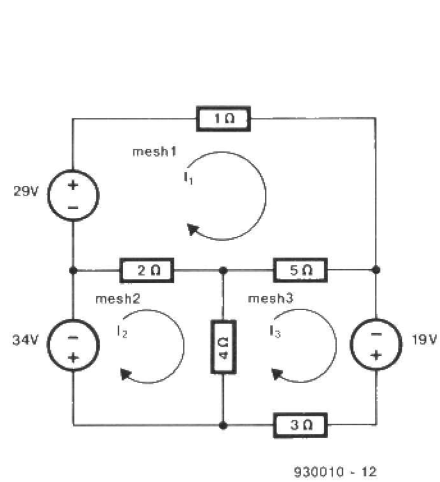


Fig. 34a. Mesh current analysis: circuit.

$$I_1 = \frac{\begin{vmatrix} 29 & -2 & -5 \\ -34 & 6 & -4 \\ 19 & -4 & 12 \end{vmatrix}}{\begin{vmatrix} 8 & -2 & -5 \\ -2 & 6 & -4 \\ -5 & -4 & 12 \end{vmatrix}} = \frac{850}{170} = 5$$

$$I_2 = \frac{\begin{vmatrix} 8 & 29 & -5 \\ -2 & -34 & -4 \\ -5 & 19 & 12 \end{vmatrix}}{\begin{vmatrix} 8 & -2 & -5 \\ -2 & 6 & -4 \\ -5 & -4 & 12 \end{vmatrix}} = \frac{-340}{170} = -2$$

$$I_3 = \frac{\begin{vmatrix} 8 & -2 & 29 \\ -2 & 6 & -34 \\ -5 & -4 & 12 \end{vmatrix}}{\begin{vmatrix} 8 & -2 & -5 \\ -2 & 6 & -4 \\ -5 & -4 & 12 \end{vmatrix}} = \frac{510}{170} = 3$$

Fig. 34b. Mesh current analysis: solution by determinants.

Knowing these currents and the resistances, we can also work out the voltages at each node. This is an exercise for the reader (solution on p.47). This network can be solved by using superposition instead—another exercise for the reader).

If there are two or more voltage sources in the mesh, these are summed, taking account of polarity.

More meshes

The network in Fig. 34a has three meshes, with three mesh currents. It gives rise to these three mesh equations:

mesh 1: $8I_1 - 2I_2 - 5I_3 = 29$;

mesh 2: $-2I_1 + 6I_2 - 4I_3 = -34$;

mesh 3: $-5I_1 - 4I_2 + 12I_3 = 19$.

The value of –34 for mesh 2 shows that the polarity of the source is such as to drive the current in the opposite direction to the specified I_2.

We now have three variables and three simultaneous equations. By calculator, using a built-in program for solving three simultaneous equations, we discover that:

$I_1 = 5$;

$I_2 = -2$;

$I_3 = 3$.

Note that I_2 is negative because it flows in the opposite direction to that indicated by the arrow. Without a programmed calculator, the equations are solved by the standard method but, with three (and even more so with four or more meshes and equations), the calculations become a tangle.

To systematize the solution of simultaneous equations of third and higher orders, two main arithmetical algorithms have been developed, which are usually featured in textbooks of electrical and electronic engineering. One of these relies on **determinants**. This method, sometimes called **Cramer's Rule**, is illustrated in Fig. 34b and we shall discuss it below. For circuits with four or more meshes, it is **possible** but not desirable to write and solve equations involving 4th-order or higher order determinants. More about solution methods in a moment.

Another technique for systematizing the solution of simultaneous equations is based on matrix arithmetic. The most commonly used method is the **augmented matrix** in which the coefficients and constants are manipulated within a matrix according to a given set of rules. While matrix arithmetic is a fascinating branch of mathematics in its own right and has applications in many field, it has only one important application in electrical and electronic engineering, which is the solution of simultaneous equations. The matrix method is probably slightly easier to use than the determinant method, but determinants lend themselves better to programming.

Determinants

Later in this article we provide a computer program in BASIC for evaluating determinants up to the fifth order. First we shall look briefly at the usual evaluation technique, partly to help the reader understand the program, and partly because mesh equations may sometimes include complex numbers, integrals or trig functions that are beyond the scope of the BASIC program. We demonstrate the technique by example, evaluating the numerator determinant for I_1 from Fig. 34b.

$$\begin{vmatrix} 29 & -2 & -5 \\ -34 & 6 & -4 \\ 19 & -4 & 12 \end{vmatrix}$$

Write out each element of the top row, multiplied by its **co-factor**:

$$+29\times\begin{vmatrix} 6 & -4 \\ -4 & 12 \end{vmatrix}$$

$$-2\times\begin{vmatrix} -34 & -4 \\ 19 & 12 \end{vmatrix}$$

$$-5\times\begin{vmatrix} -34 & 6 \\ 19 & -4 \end{vmatrix}$$

$$\begin{vmatrix} + & - & + & - & + & \cdots \\ - & + & - & + & - & \cdots \\ + & - & + & - & + & \cdots \\ - & + & - & + & - & \cdots \\ \cdots & & & & & \end{vmatrix}$$

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Fig. 35. Place signs of co-factors.

The co-factor of an element is the determinant written out again, but omitting the row and the column in which the element occurs. Thus, the co-factors of elements of a third-order determinant are all second-order determinants. Prefix each product with a plus or minus sign according to the position of the element. Figure 35 shows the place sign for the co-factor of each element; in the example, the signs are +, –, +, so that:

$$+29\times\begin{vmatrix} 6 & -4 \\ -4 & 12 \end{vmatrix}$$

$$-(-2)\times\begin{vmatrix} -34 & -4 \\ 19 & 12 \end{vmatrix}$$

$$+(-5)\times\begin{vmatrix} -34 & 6 \\ 19 & -4 \end{vmatrix}$$

Now to evaluate these second-order determinants. The rule is the same as for third-order determinants. Taking the first determinant as an example, write out each element of the top row, multiplied by its co-factor and prefix each product by the appropriate place sign:

$$\begin{vmatrix} 6 & -4 \\ -4 & 12 \end{vmatrix} = +6\times12-(-4)\times(-4)$$
$$=72-16=56.$$

Here, the co-factors are the single elements 12 and –4, so the products can be calculated and summed without further ado. The rule for evaluating second-order determinants may be more conveniently expressed as **the product of the elements on the leading diagonal minus the product of the elements on the other diagonal**. In a similar way, we find that the values of the other two determinants are:

$$(-34)\times12-(-4)\times19 =$$
$$= -408-(-76) = -332;$$

and

$$(-34)\times(-4)-6\times19 =$$
$$= 136-114 = 22.$$

We are now ready to substitute these values into the equation for the value of the original third-order determinan:

$$29\times56-(-2)\times(-332)+(-5)\times22=$$
$$=1624-664-110=850.$$

In the same way, we find the values of the other determinants listed in Fig. 34b and then calculate the currents.

We evaluated the determinant by taking the elements from the top row together with their co-factors. The same result is obtained if we take any other row instead, or any column. This fact may be made use of where a determinant contains any zeros. By basing the calculations on a row or column containing one or more zeros, some of the elements are zero and the number of products to be summed is correspondingly reduced.

Higher orders

Evaluating a third-order determinant is lengthy enough, with ample opportunity for arithmetical errors. Evaluating a fourth-order determinant is even more horrendous. The original determinant is written out as the four elements of a row or column, each multiplied by its third-order co-factor. Each of these four third-order co-factors has to be evaluated in the same way as third-order determinants, giving altogether 12 second-order determinants to be evaluated. As the

number of meshes in the circuit increases, the computation becomes more and more unwieldy. This is why the BASIC program in Fig. 36 is so useful.

The pogram is written in LOCOMOTIVE BASIC, but we have restricted it to its bare bones so that it is readily convertible into other BASIC dialects, such as GW/BASIC. We leave it to the reader to refine the program by adding error-trapping routines to the input lines and by formatting the display to suit the machine.

The program evaluates determinants of any order up to five. It is easy to add further sub-routines using the same algorithm to cater for determinants of the sixth, seventh, or even higher orders. It is also feasible to change the input routine and extend the program so that it accepts the coefficients of all the mesh equations at one time, calculates the required determinants and then directly calculates the currents.

The program consists mainly of a set of nested sub-routines. At the heart of the nest, a second-order determinant is evaluated according to the 'leading diagonal minus other diagonal' stated above (sub-routine 200). The co-factors of a third-order determinant are assembled by sub-routine 300. Each of these is evaluated in turn by repeatedly calling sub-routine 200. In sequence, these are multiplied by the appropriate element and sign. The program takes the elements from the top row of the determinant. A fourth-order determinant is handled by sub-routine 500, which calls sub-routine 300 to evaluate the third-order co-factors, which calls sub-routine 200 for the eventual evaluation of the second-order co-factors. A fifth-order determinant is evaluated by sub-routine 700, which calls sub-routine 500 and ultimately sub-routines 300 and 200.

To asist the reader in expanding or adapting the program, here is a variable and array list:

b3, b4, b5	The number of the column of the co-factor currently being assembled.
c3, c4, c5	The number of the column of 3rd, 4th or 5th order determinants currently being assembled into a co-factor (one of these columns is omitted from the co-factor, the column in which the current element occurs).
j, k	Loop counters.
m3, m4, m5	The number of the column of 3rd, 4th or 5th order determinants, used to specify for which element of the top row the co-factor is currently being assembled.
n	Order of the determinant being evaluated.
si3, si4, si5	+1 or −1, the sign for each element/co-factor product in 3rd-, 4th- and 5th-order sub-routines,
v, w, x, y	Output from sub-routines, sums of products of each element and its co-factor; for exmple, returning from sub-routine 200, v holds the value of the 2nd-order determinant; this is multiplied by the element and sign; successive values of v are accumulated in w (see line 370); similarly, in a 4th-order evaluation, successive values of w are accumulated in x (see line 570).
z	Final value of determinant.
d2(), d3(), d4(), d5()	Arrays to hold determinants of orders 2 to 5 (these are not DIMmed in the program, as that is unnecessary in this version of BASIC; their dimensions are 2×2, 3×3, 4×4 and 5×5 respectively).

Answers to Test Yourself – Part 3

1. 2.2 A
2. 3.5 A
3. $I=0.6$ A; $U=4$ V; no effect
4. $I=1.125$ A; $U=-1.75$ V
5. $U_{TH}=2$ V; $R_{TH}=3.2\ \Omega$
6. $U_{TH}=0.8$ V; $R_{TH}=1.2\ \Omega$

```
10 REM: DETERMINANTS ***
20 CLS:n=0
30 WHILE n<2 OR n>5
40 INPUT"Order";n
50 WEND
60 PRINT
70 FOR j=1 TO n
80 PRINT "Row ";j
90 FOR k=1 TO n
100 PRINT "    Column ";k
110 INPUT d5(j,k)
120 NEXT:NEXT
130 GOSUB 900
140 ON n GOSUB 10,200,300,500,700
150 PRINT:PRINT"Value = ";z
160 FOR j=1 TO 5: PRINT:NEXT
170 PRINT"Any key for another"
180 WHILE INKEY$="":WEND
190 GOTO 20
200 IF n=2 THEN FOR j=1 TO 2:FOR k=
1 TO 2:d2(j,k)=d5(j,k):NEXT:NEXT:RE
M SECOND ORDER ***
210 v=d2(1,1)*d2(2,2)-d2(1,2)*d2(2,
1)
220 IF n=2 THEN z=v
230 RETURN
300 IF n=3 THEN FOR j=1 TO 3:FOR k=
1 TO 3:d3(j,k)=d5(j,k):NEXT:NEXT:RE
M THIRD ORDER ***
310 v=0:w=0:si3=1
320 FOR m3=1 TO 3:b3=1
330 FOR c3=1 TO 3
340 IF c3<>m3 THEN d2(1,b3)=d3(2,c3
):d2(2,b3)=d3(3,c3):b3=b3+1
350 NEXT
360 GOSUB 200
370 v=v*d3(1,m3)*si3:w=w+v:si3=-si3
380 NEXT
390 IF n=3 THEN z=w
400 RETURN
500 IF n=4 THEN FOR j=1 TO 4:FOR k=
1 TO 4:d4(j,k)=d5(j,k):NEXT:NEXT:RE
M FOURTH ORDER ***
510 x=0:si4=1
520 FOR m4=1 TO 4:b4=1
530 FOR c4=1 TO 4
540 IF c4<>m4 THEN d3(1,b4)=d4(2,c4
):d3(2,b4)=d4(3,c4):d3(3,b4)=d4(4,c
4):b4=b4+1
550 NEXT
560 GOSUB 300
570 w=w*d4(1,m4)*si4:x=x+w:si4=-si4
580 NEXT
590 IF n=4 THEN z=x
600 RETURN
700 y=0:si5=1:REM FIFTH ORDER ***
710 FOR m5=1 TO 5:b5=1
720 FOR c5=1 TO 5
730 IF c5<>m5 THEN d4(1,b5)=d5(2,c5
):d4(2,b5)=d5(3,c5):d4(3,b5)=d5(4,c
5):d4(4,b5)=d5(5,c5):b5=b5+1
740 NEXT
750 GOSUB 500
760 x=x*d5(1,m5)*si5:y=y+x:si5=-si5
770 NEXT
780 IF n=5 THEN z=y
790 RETURN
900 CLS:FOR j=1 TO n:REM PRINT DET
***
910 FOR k=1 TO n
920 PRINT d5(j,k);"    ";
930 NEXT:PRINT:NEXT
940 PRINT:PRINT
950 RETURN
```

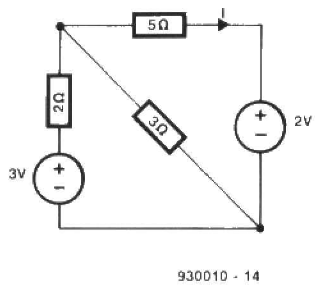
Fig. 36. BASIC listing.

Test yourself

For each of the circuits in Fig. 37, use the mesh current method to find *I* to three decimal places or as fractions. Confirm your result by using superposition.

Answers will be given in next month's instalment.

a



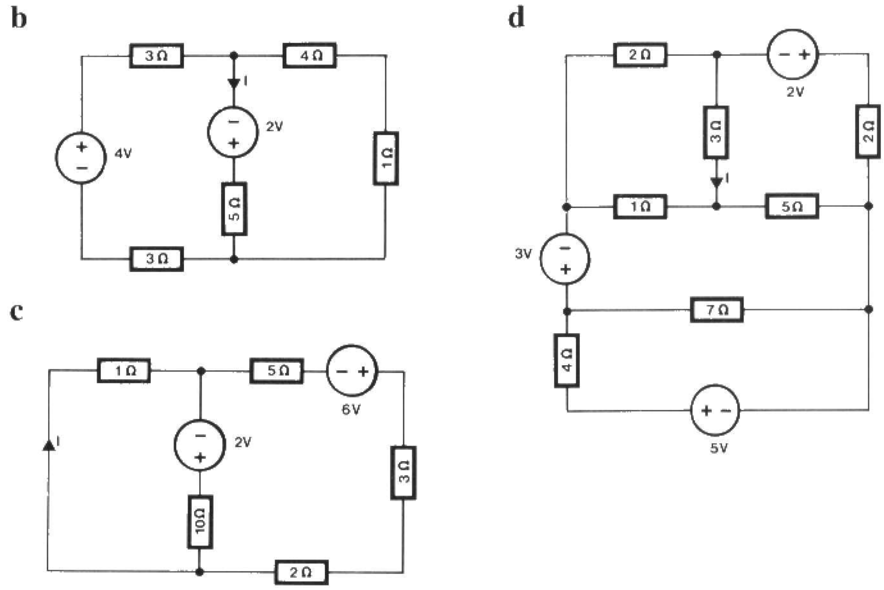


Fig. 32. Networks for Test Yourself.

SAGE 'SUPER - LINK ' PREAMP

The perfect link

To complement our new range of class A power amplifers we now launch a complete class A preamplifier the **SAGE** 'Super-link' system. The preamp is in modular form comprising four separate stereo modules all fully assembled and tested. Assembly is thus straightforward consisting of simple mounting and wiring to produce a top class matched preamp. All modules can be used independently or as a whole preamp system, the perfect link to the **SAGE** Supermos modules

Module 1 The 'Selector' A complete, ready built stereo signal source selector, features, gold phono stereo in/outs source selection without mechanical or electronic switches in the signal path, available with or without phono equaliser stage.

Module 2 The 'Controller' A complete ready built stereo control module features pure class A operation,total control over volume, bass, treble, balance, active adjustable gain stage

Module 3 The 'Power Supply' The most perfect power supply avaiable, virtual zero output impedance, noise and total absence of ripple, powers all four modules

Module 4 The 'Equaliser' An optional plinth mountable class A phono equaliser amplifier for both MC and MM cartridges, adjustable gain,superior performance.

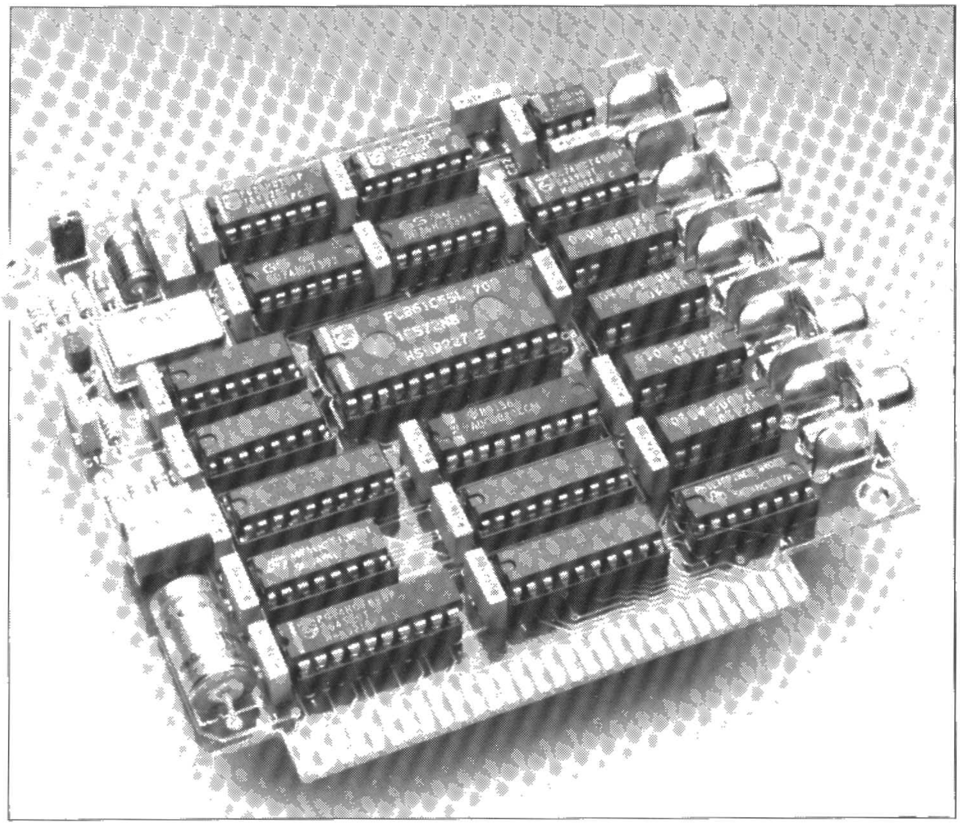
NEW PRODUCTS FOR 92/93

Sage Audio have launched many new and exciting produck for 92/93, these include a complete new range of class A power amplifier modules the **Supermos200 to Supermos1000** range with absolutely exemplary perfofmance unmatched anywhere in the audio industry.

Specs include maximum power output range from 50 watts to 1000W, THD less than 0.0001%, slewrate over 700V/us, ripple rejection virtual infinite, freq resp' 0.5Hz - 350Hz and there's much more we could fit in this ad.

For full details of our full range of class A audio products full amplifier kits mains filters and price lists please send a **large stamped addressed envelope and £2 coins** to:-
(Overseas please send 6 International reply coupons)
SAGE AUDIO Construction House, Whitley St, Bingiey, Yorks, BD16 4JH, England.

VIDEO DIGITIZER FOR PCs



The insertion card described here allows you to put TV images on your computer screen. Designed to handle colour images at a resolution of 24 bits, the card is easy to build, and forms a perfect link with several graphics workshop programs.

Design by John Kortink

THE call for more and better tools to integrate text, images and sound on a PC has had a new impulse with the widespread introduction of the Microsoft Windows graphics user interface (GUI). An advantage of a GUI is that the screen shows exactly what will appear on paper. This used to be impossible with text-only systems, which were only capable of giving a rough indication of the final output. Furthermore, there is a lot of activity on the printer market. Inkjet printers that produce good colour prints are now available at affordable prices. An example is the Hewlett-Packard Deskjet 500C. The colour prints shown on this month's front cover were made using a satellite TV receiver as the video source, the video digitizer, Corel Draw 3.0, and a Canon BJC800 colour printer.

Since the Windows GUI offers excellent colour support, and colour printers are now in the shops, we decided to make the present video digitizer compatible with 24-bit colour files. The software supplied with the digitizer board packs the digital colour information into a TIFF (Tagged Image File Format) file, which can be processed further by almost any graphics program that runs under MS-DOS or Windows. The DOS program supplied with the digitizer board only reads the digitized video signals, and translates this information into a TIFF file. The TIFF format supports several sub-formats: 1, 8 and 24 bits, black-and-white and colour. Those of you who do not require 24-bit colour resolution may also use the digitizer to produce images with 256 levels of grey. Everything required to achieve this is offered by the hardware and the software described here.

The digitizer is a relatively simple type that requires the input image to be still for some time — the minimum 'still' time is about 2.5 s for a black-and-white image, and about 15 s for a colour image. Most modern video recorders and camcorders are capable of producing quite acceptable still images. All-solid-state still-video recorders like the Canon ION obviously have no trouble at all providing still images. The composite video signal supplied by, for instance, a video recorder may be used to capture black-and-white images. The digitizer has separate RGB (red-green-blue) inputs for colour images. To convert from composite colour (CVBS) into RGB, we suggest the use of the S-VHS/CVBS-to RGB converter described in Ref. 1.

The circuit

Digitizing a video image is pretty complex, and the timing of the whole process is critical. In addition, we have to take into account the huge speed differences between PCs — an old XT runs at snail's pace compared to a 486! This explains the presence of a buffer in the digitizer. This buffer enables the PC to process the video information at its own speed.

Looking at the circuit diagram, Fig. 1, you may find that this is very similar to the earlier video digitizer for the Archimedes computer (Ref. 2). The optional colour extension for the older design is now incorporated in the circuit. Although we are looking at a pretty complex circuit, it was still possible to accommodate it on a 'short' PC insertion card.

The PC extension address bus interface is formed by IC_1 (74HCT688) and IC_2 (74HCT138). These two ICs arrange the address decoding of the card. The card address is set with the aid of three DIP switches (S_1). The basic card address used by the software is \$300 (all switches closed). In the PC's I/O address space, location \$300 is reserved for experimental insertion cards, and should be free in most cases. In the rare case of a different address being required, the DIP switches come to your aid. Also, the variable 'dib' has to be changed accordingly in the source file, and the program has to be recompiled (Turbo Pascal 6.0).

The four selection signals needed by the circuit are generated by demultiplexer IC_2 from the address decoder enable signal (pin 19 of IC_1), address line A0, IOR and IOW. The next three integrated circuits, IC_3, IC_4 and IC_5, ensure the necessary buffering be-

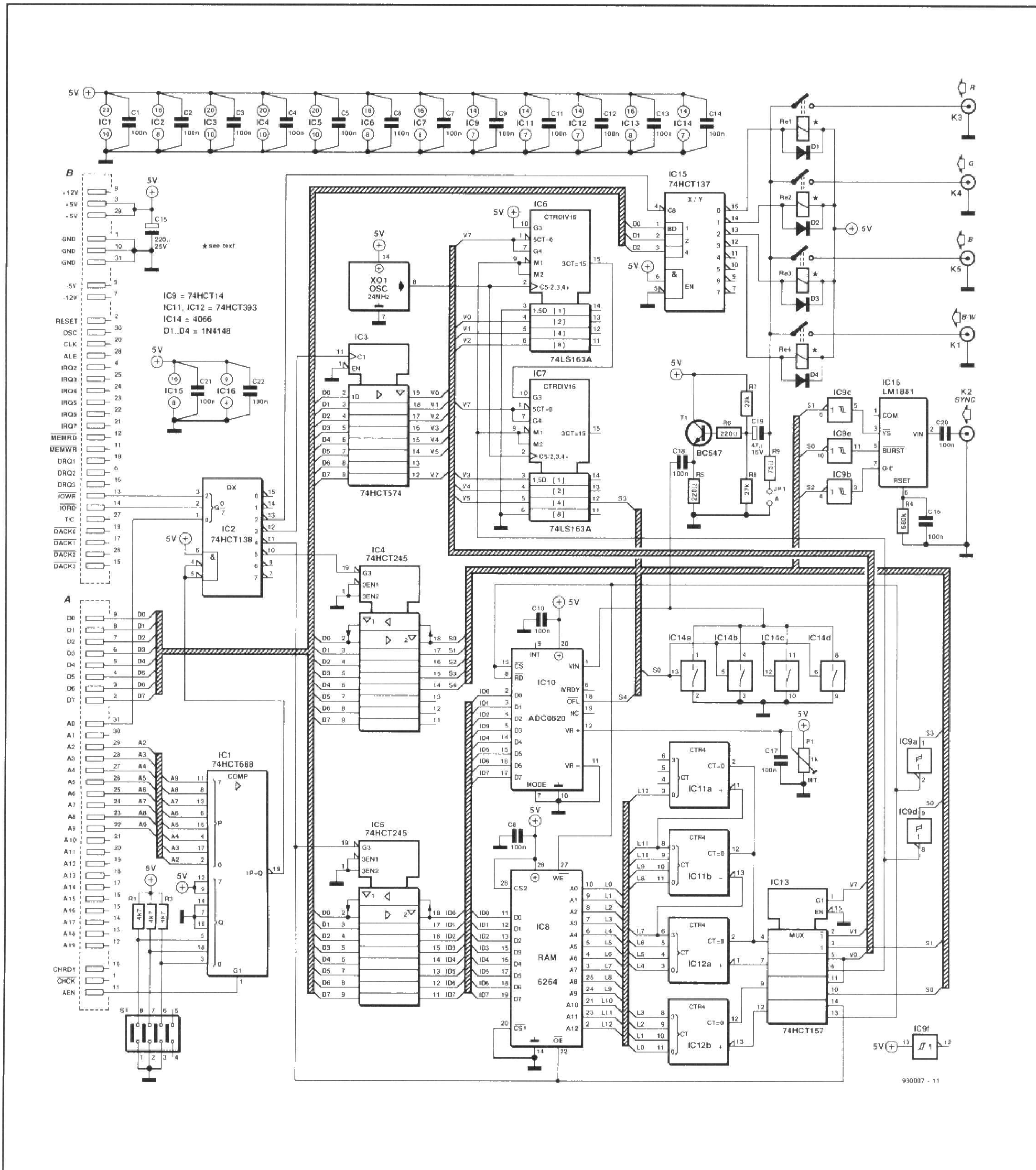


Fig. 1. Circuit diagram of the 24-bit video digitizer for MSDOS PCs. Monochrome images are digitized at 256 grey levels.

tween the rest of the digitizer and the PC's databus. The digitizer uses a databus width of 8 bits, which allows even PC-XTs to make use of the hardware.

The right-hand side of the diagram indicates how the video signals arrive in the circuit. Basically, there are two signals: the video signal (RGB or monochrome, switched by a relay), and the synchronization signal, SYNC. The video signal is fed to the A-D converter Type ADC0820 (IC10) via a buffer transistor, T1. Four electronic switches, IC4a-IC4d, are used to make sure that there is no signal at the ADC input during the dead time of the video. The reference signal required by the ADC to digitize the video information is available at pin 12 (VR+) of IC10. Preset P1 is used to optimize the conversion by setting the correct black and white levels.

The ADC input signal is selected with the aid of four relays. Monochrome signals are selected by actuating relay Re4 via IC15. This results in the signal at the B/W input (K1) being passed to the ADC. When colour signals are digitized, the rasters at the R, G and B inputs (K3, K4, K5) are successively applied to the ADC input. After the three colour signals have been sep-

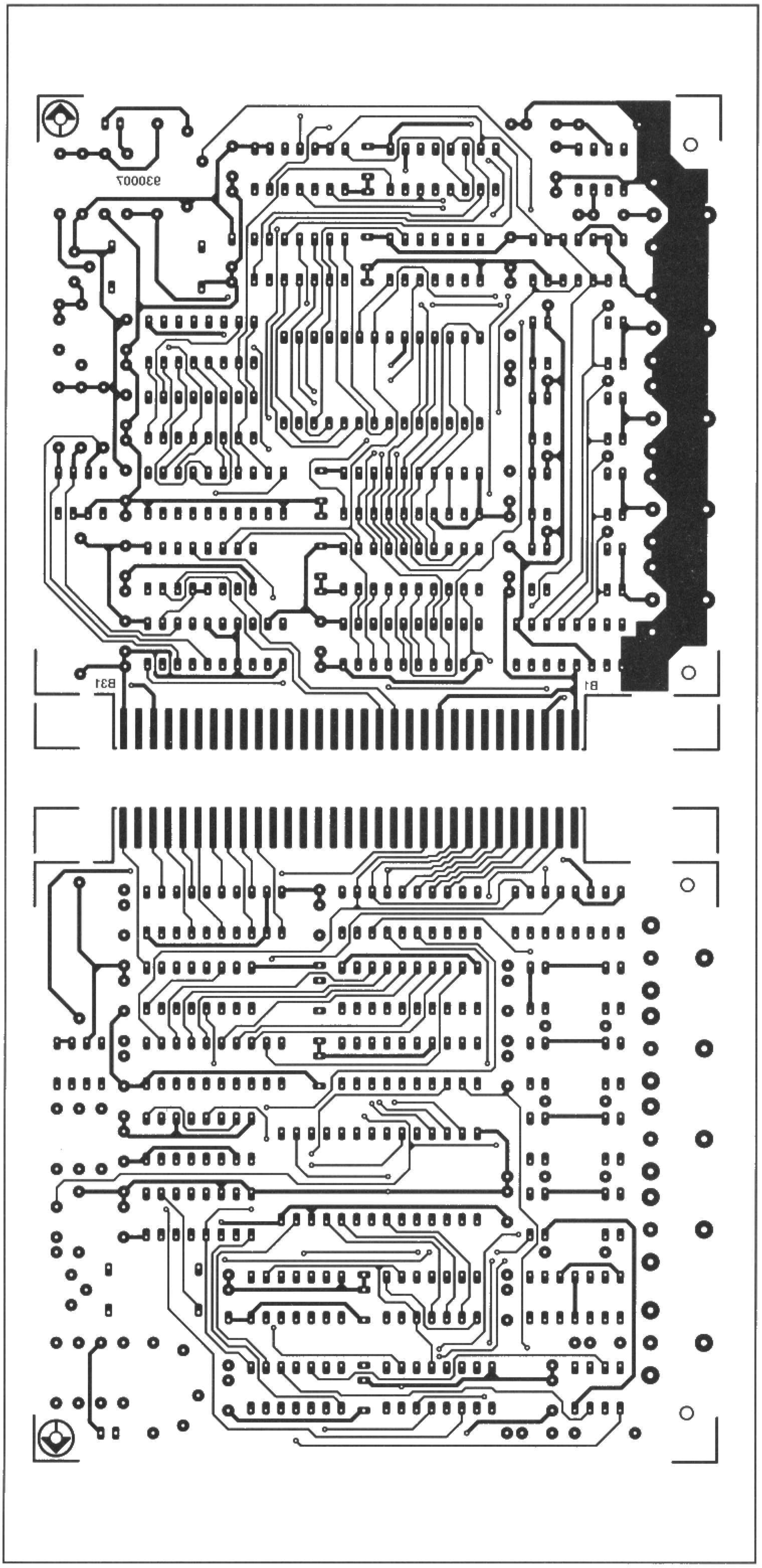


Fig. 2a. Track layouts of the double-sided through-plated PCB designed for the video digitizer.

arately digitized (this is done in exactly the same way as with a black-and-white file), and have been stored in three temporary files, the software puts the information together into one large TIFF file.

The other input, K2, carries the SYNC signal, which is 'dissected' into three components by IC16, an LM1881: BURST (pin 5), vertical synchronization (VS, pin 3) and Odd/Even raster (pin 7). The digitizer uses the BURST signal as the horizontal sync pulse.

The horizontal sync signal has an important function in the digitizer hardware, while the vertical sync signal is used to reset some counters. It is also used by the software, in combination with the odd/even raster signal. VS and odd/even are fed to the computer's databus via buffers contained in IC4, a 74HCT245. The electronic switches in IC14 ensure that the black level of the video signal is restored during each horizontal sync pulse.

Measuring in small steps

A complete picture line has a length of 64 μs including the synchronization pulse. The distance between successive samples is 0.1 μs at the maximum resolution of 640 pixels. This means that the ADC used is far too slow since it achieves a minimum conversion time of 1.2 μs. The solution to this problem is to sample each picture line several times to obtain the desired information. Apart from the hardware, there are further limiting factors, including the software and the data storage in the 8-KByte cache. All in all, it is required to sample each picture line 64 times before it is completely digitized. This results in a sample interval of 6.4 μs, and, consequently, a minimum 'still' time of $\frac{1}{25} \times 64 = 2.56$ s (remember, 25 odd and 25 even ras-ters are transmitted every second).

Since every picture line has to be sampled so many times, the sample timing needs to be accurately defined. This is achieved by dividing the 6.4-μs interval into 64 slots of 0.1 μs each. The sample instant shifts exactly one slot each time the line is sampled. This eventually produces the complete information on the picture line in the memory. After the digitizer software has done its bit on the collected data, the 640 samples are arranged in the right order again.

The digitizing operation

A video signal is digitized by taking samples of its brightness (luminance) component(s). This is achieved with the aid of IC10. A sample is taken on the start signal (pins 8 and 13 are made low simultaneously, while pin 7

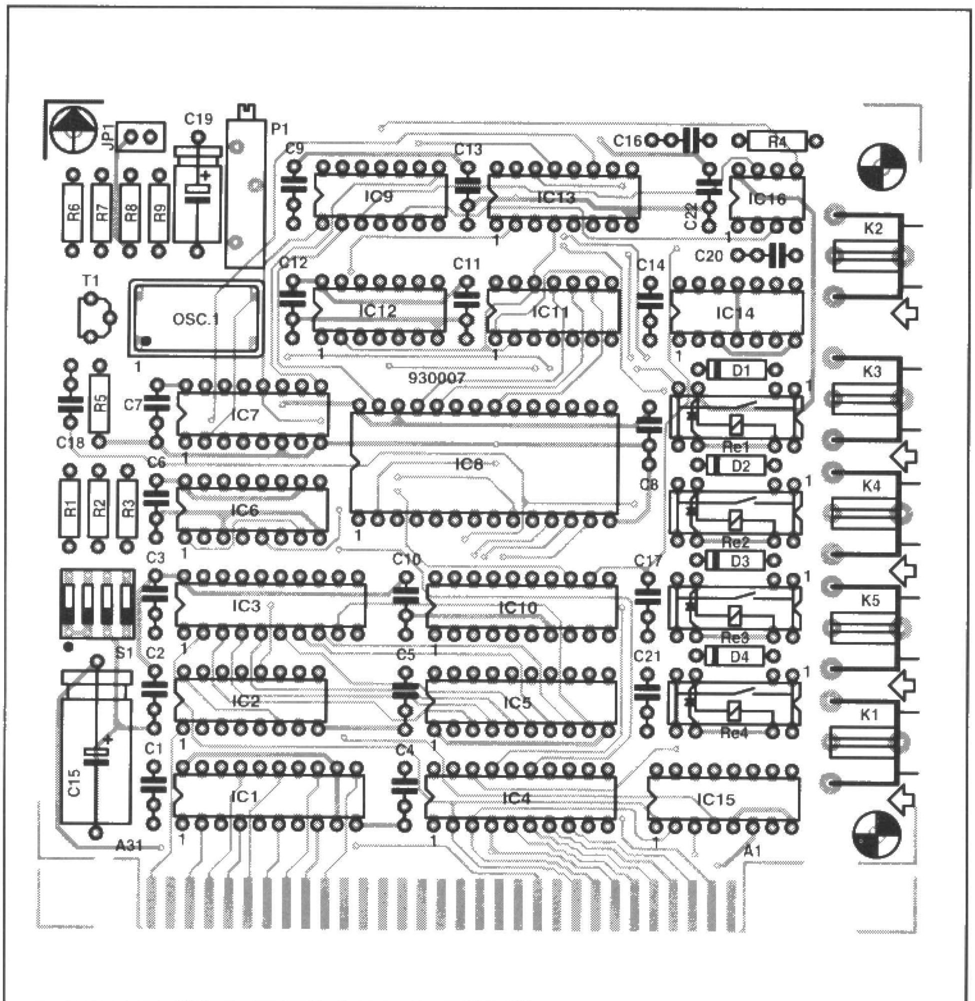


Fig. 2b. Component mounting plan.

is low already). The sample is stored internally, and takes about 1.2 µs to convert into a digital value, which is processed further by the PC. Although it would have been possible to have the PC read the data during a DMA operation, this is not done here to ensure that the digitizer hardware can be used on all PCs (DMA requests, timing and servicing are notorious sources of compatibility problems in the PC world). All samples generated while the picture is being digitized are temporarily stored in an 8-KByte cache memory, IC8. Each line is sampled ten times for each digitizing operation, so that, at the highest resolution, each block consists of ten samples times 512 lines. That is why the digitizing instruction is repeated several times, every iteration starting at a different off-set with respect to the start of the picture line.

The cache memory is associated with a 13-bit address counter built around IC11a, IC11b, IC12a and IC12b. The start-of-conversion pulse (output Qc, pin 12 of IC7) not only triggers the ADC, but also serves to put a write pulse on the WE\ (write enable) input of the memory. Since the RAM copies the data on the databus on the positive edge of this pulse, the data that has been available on the ADC outputs can be copied without problems.

There is another stumbling block

COMPONENTS LIST

Resistors:

3	4kΩ7	R1;R2;R3
1	680kΩ	R4
2	220Ω	R5;R6
1	22kΩ	R7
1	27kΩ	R8
1	75Ω	R9
1	1kΩ multiturn preset	P1

Capacitors:

20	100nF	C1-C14;C16; C17;C18;C20; C21;C22
1	220µF 25V	C15
1	47µF 16V	C19

Semiconductors:

4	1N4148	D1-D4
1	BC547	T1
1	74HCT688	IC1
1	74HCT138	IC2
1	74HCT574	IC3
2	74HCT245	IC4;IC5
2	74LS163A	IC6;IC7
1	6264 or 61C65-70n	IC8
1	74HCT14	IC9
1	ADC820CCN*	IC10
2	74HCT393	IC11;IC12
1	74HCT157	IC13
1	74HCT4066	IC14
1	74HCT137	IC15
1	LM1881*	IC16

Miscellaneous:

1	Jumper	JP1
5	PCB mount cinch socket	K1-K5
1	24MHz oscillator block	XO1
4	V23100-A4005-A010** relay	Re1-Re4
1	Insertion card mounting bracket	
1	PCB and software on disk (1831); order code 930007 (see page 70)	
1	Software on disk; order code 1831 (contained in set 930007) (see page 70)	

* National Semiconductor product. National Semiconductor, The Maple, Kembrey Park, Swindon SN2 6UT. Tel.: (0793) 697428; fax (0793) 697706. Distributors: Abacus Electronics Ltd., tel.: (0635) 33311, fax: (0635) 38670; ESD Distribution Ltd., tel.: (0279) 626777, fax.: (0279) 441687; Farnell Electronic Components Ltd., tel.: (0532) 636311, fax.: (0532) 633411; Jermyn Distribution, tel.: (0732) 740100, fax.: (0732) 451251; Makro Marketing Ltd., tel.: (0628) 604422, fax.: (0628) 666873; Thame Components Ltd., tel.: (0844) 261188, fax.: (0844) 261681.

** Siemens product. ElectroValue, 28 St Jude's Road, Englefield Green, Egham, Surrey TW20 0HB. Telephone: (0784) 433603. Fax: (0784) 435216. C-I Electronics, P.O. Box 22089, 6360 AB, Nuth, Holland. Fax: +31 45 241877.

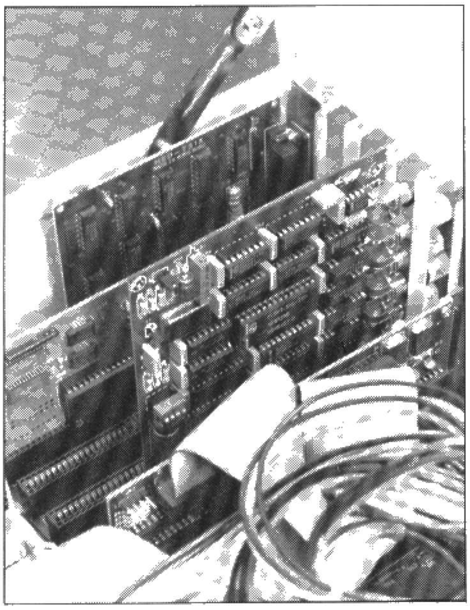


Fig. 3. Prototype of the digitizer fitted into a PC-AT. The contrast preset may be moved from the PCB to the fixing bracket, but only if there is sufficient space (for instance, if you have omitted some of the input sockets). A readily accessible contrast control is particularly useful if you are using many different video sources.

that has to be overcome: the distance between two successive pulses is only fixed within a picture line — at a transition between two picture lines, it is necessary to wait, for instance, on the burst pulse. This problem is solved by dividing the counter into a samples-per-line counter and a line counter. To keep the samples counter as simple as possible, the lower four bits of the address counter are assigned to the samples-per-line counter (IC_{12b}). This reserves enough memory for this part of the operation. The remaining 9 bits are used by the line counter.

The reset and clock lines of the line counter are controlled via the four outputs of IC_{13}. This multiplexer selects between two totally different drive signal combinations, depending on the level at bit 7 of buffer IC_3 (pin 12). During the digitizing operation, bit 7 of IC_3 is high, while it is low during reading. While bit 7 is high, the contents of the line counter is increased by one on every burst pulse. The resetting is done by the vertical sync pulse, VS, which, up to the end of its duration, keeps the counter state at nought, in spite of the horizontal sync pulses that are generated during this period.

The sample counter is advanced on every negative edge of the start-of-conversion pulse. This means that the counter is increased before the first byte is stored, and, consequently, that the first byte is not used. This is not a problem, however, since there are 16 locations available for the ten samples taken per picture line. The sample counter is reset by the burst signal, i.e., at the same time the line counter is increased.

During the read operation (bit 7 of IC_3 low), the 8-bit counter formed by IC_6 and IC_7 is reset and disabled. Since the inputs for parallel loading (pin 1) and resetting (pin 1) are both low, the counter outputs cannot change, and remain at nought. At the same time, the ADC is blocked (via inverter IC_{9a}) and also does not supply signals to the system. This 'freezes' the digitizer, allowing the software to read the data stored in the RAM chip during the digitizing operation.

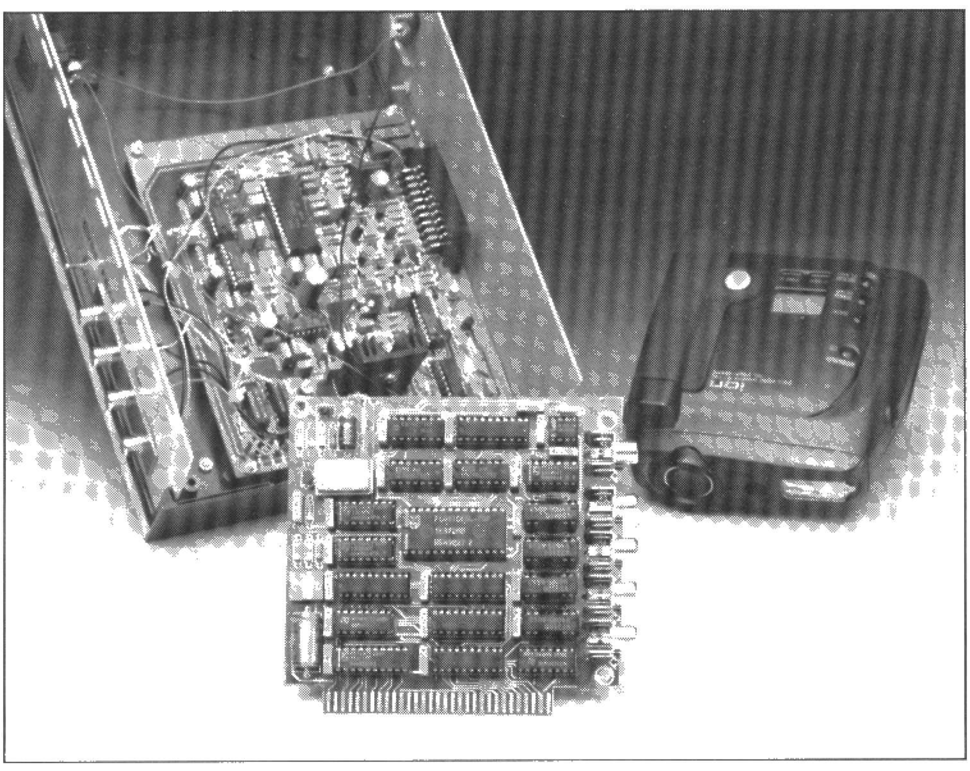
Next, the line counters are controlled by the software, via bits 0 and 1 of IC_3. The sample counter is reset every time the line counter is increased via bit 0 of IC_3. Next, it is increased on every read pulse. In this way, each location in the RAM memory is read out.

A matter of counting

The exact starting point of a series of samples is determined by IC_6, IC_7 and IC_{13}. The 8-bit counter formed by IC_6 and IC_7 is clocked by a 24-MHz central clock signal supplied by an integrated quartz oscillator block.

As the counter states cycle, outputs QA-QD (pins 11-14) supply period times of 1.67 µs, 5.33 µs and 10.67 µs. The digitizer uses the 5.33-µs signal to control the ADC. This means a deviation of 1.07 µs from the ideal period time of 6.4 µs. In practice, this is not a problem — the start and the end of a line need not be sampled because of the sync signals. Consequently, a period of 53.3 µs of the total video line time of 64 µs is sampled.

The 8-bit counter is loaded with the contents of IC_3 on every horizontal sync pulse. This value determines the time between the sync pulse and the first 'low' on the Q_C output of IC_7. From then on, this output will go low again after 5.33-µs pauses, causing the ADC to take a sample. After the start point has been shifted a little 64 times in this way, the contents of the whole line has been sampled.



Construction and test

The digitizer is built on a double-sided, through-plated printed circuit board with gold-plated contact fingers for insertion in a PC extension bus slot connector. The card is secured to the back panel of the PC by a cover bracket, which is drilled to pass the five 'phono' sockets. The cover bracket need not be purchased — simply remove and drill the one at the location you wish to insert the digitizer card. Those of you who do not require the colour inputs may simply omit the relevant PCB sockets. It is even possible to reduce the number of sockets to one if monochrome images are applied in the form of a CVBS (composite video) signal. In that case, simply interconnect the CVBS and SYNC inputs.

The construction of the board is not expected to cause difficulty. The ICs may be fitted in sockets. Jumper JP_1 is fitted only if it is required to terminate the video source into 75 Ω. This will be necessary with most video sources —

remember, mismatches can give rise to complex faults.

A couple of experiments are required to adjust preset P1 for the highest possible brightness of the digitized pictures. The simplest way to adjust the preset is to digitize a TV test chart, and evaluate the results. If an S-VHS source is used to supply monochrome images, the picture quality can be improved significantly by using the luminance component only. This eliminates moiré effects in the picture as a result of cross-interference between the luminance and chrominance components in a CVBS signal.

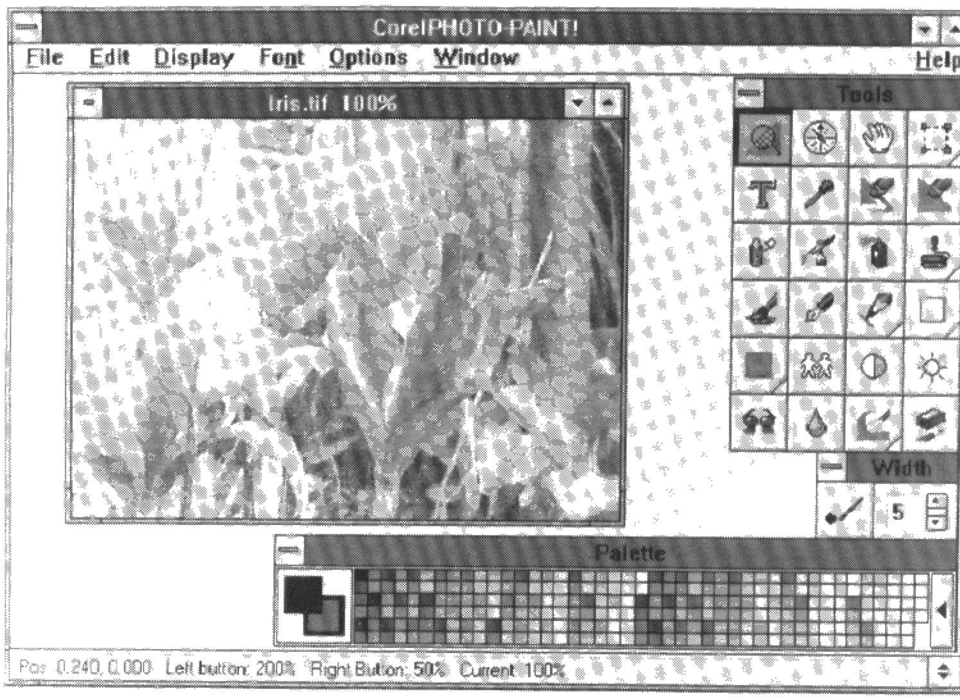


Fig. 4. Corel PhotoPaint! screendump showing an imported digitized picture (TIFF format) ready for editing.

On control software

The operation of the video digitizer hardware is supported by a control program. The Turbo Pascal V 6.0 source code of this program is also supplied on the project diskette, which may be obtained through our Readers Services as order code 1831. The source code is useful for those of you who wish to extend the program, or adapt it for use under Windows 3.1.

The picture digitizing operation is started with the aid of the PCDIGI command and a number of software switches. The command PCDIGI /? produces a help screen — see Fig. 5. The 14 software switches shown allow you to determine what the digitized picture will look like. Switches 't', 'b', 'l' and 'r' enable digitized pictures to be trimmed. The 't' and 'b' switches cause non-used picture lines at the top and the bottom of the pictures to be cut off, while the 'l' and 'r' switches do the same with the start and the end of the picture line. The interlace switch, 'i', allows both rasters in the image to be digitized. This causes the vertical resolution to increase to 640 pixels. Alternatively, switches 'o' and 'e' may be used to select the odd and even raster respectively. Switch 's' is used to select the video signal to be digitized (R, G, B or CVBS). This switch will rarely be used because the software controls the switches itself when digitizing colour or monochrome images. Finally, the 'q' (quiet) and 'v' (verbose) switches determine whether or not status messages appear on the screen during the digitizing operation.

If you type PCDIGI <filename> without any switches, the program digitizes the odd raster in black-and-white, without trimming the sides of the image. The screen displays process reports. The variable filename may be preceded by a path indication. To ensure the highest possible speed, it is recommended to use paths to files and directories on the hard disk only.

Apart from the executable code and the source code of PCDIGI, the project diskette also contains Graphic Workshop in compressed form. This program by the Canadian company Alchemy Mindworks provides many ways of viewing and editing picture formats (including TIFF). Graphic Workshop will unpack itself if you type GRAFWK61. It should be noted that Graphic Workshop is a shareware program that may be used for a limited period only if do not pay for it. If you like the program and wish to keep using it, you are asked to transfer $40.00 to the authors. In return, you will be sent the latest version of the program. Further information on this is given in the documentation file that comes with the program. In addition, there is a file that allows you to print an order form. Having worked with it for some time we are convinced that Graphic Workshop is worth the $40.00 investment! ■

```
PCDigi 1.00 (11 Nov 1992), (c)1992 Zeridajh, by John Kortink
Digitises images using the Zeridajh PC video digitiser

Usage : PCDigi [{-|/}<switch>...] <filename>

<Switches>
  ?: Display this message
  n: Non-interlaced (default, odd or even frame only, 640x320)
  i: Interlaced (both odd and even frames, 640x640)
  o: Odd frame (default, ignored for -i)
  e: Even frame (ignored for -i)
  v: Verbose, give messages (default)
  q: Quiet, no messages
  g: Digitise greyscale image (default, 1 pass)
  c: Digitise 24-bit RGB colour image (3 passes), -s ignored
  t<val>: Skip top <val> lines in output image (default 0)
  b<val>: Skip bottom <val> lines in output image (default 0)
  l<val>: Skip left <val> pixels in output image (default 0)
  r<val>: Skip right <val> pixels in output image (default 0)
  s<val>: Digitise signal <val> (0=R,1=G,2=B,3=B/W, default 3)
                                        930007-13
```

Fig. 5. Help screen produced by PCDIGI.

References:

1. 'S-VHS/CVBS-to-RGB converter', *Elektor Electronics* September and October 1990.
2. 'Black-and-white video digitizer', *Elektor Electronics* July/August 1991.

Wideband wireless data systems

By Brian P. McArdle

(1) Introduction

Radiocommunications equipment for PMR or ISM applications is designed for narrowband channels. In the majority of European countries this means a channel spacing of 12.5 kHz or 25 kHz, where the carrier frequency of a channel is 12.5 kHz or 25 kHz away from the carrier frequencies on the upper and lower adjacent channels. The method of modulation is generally FM and the maximum frequency deviation is 2.5 kHz or 5.0 kHz respectively. In order to confine transmissions to specific narrow channels, spurious emissions outside an assigned channel are generally required to be at least 55 dB below the level of the nominal carrier frequency. Hence, the term narrowband FM is often used to describe the modulation method. This is in contrast to broadcasting (e.g., sound broadcasting in Band 2) where the frequency deviation is 75 kHz. Most low cost scanners distinguish between wideband FM for broadcasting and narrowband for business applications. When data communications became necessary, the techniques of modulation were adapted to allow for digital signal processing. For example, 2-tone FSK uses two different frequencies to represent the bits '1' and '0'. The UK Performance Specification MPT1317 (Appendix 1) is a protocol for data transmission at 1200 baud in the PMR service. Recently, the question of the use of wideband modulation techniques has arisen.

ETSI has approved a standard on Wideband Wireless Data Systems (WWDS) for public enquiry. It is a specification for ISM equipment in the band 2.4–2.5 GHz and is really a radio-lan. (Further standards in the general area of radio-lans will be developed, but specific areas will probably be covered by their own individual specifications). It differs from other standards for radiocommunications equipment in that spread spectrum modulation is specified. The emissions are deliberately spread over a much larger part of the spectrum than would normally be used. Two main methods of spread spectrum modulation are specified: Direct Sequence (Section 4) and Frequency Hopping (Section 5). The purpose of this article is to provide an overview of this new development. The comments are purely personal and the reader is referred to the specification for a detailed explanation.

(2) Spread spectrum modulation

This form of modulation results in a transmission with a bandwidth much wider than would be required by the original message (i.e., data). In addition, the bandwidth should not be determined by the message but by a prescribed modulating signal (baseband). Wideband FM is not classified as spread spectrum because it does not satisfy the second condition. The ratio of the bandwidth following the spread spectrum operation to the bandwidth required by the message is known as the **process gain**. This could be 30 dB or higher depending on the system. For example, a signal which originally required 3 kHz would be spread over 3000 kHz.

Since a transmission is intentionally spread out, it is difficult to detect unless a receiver can reproduce the exact process. Consequently, it is difficult to jam either deliberately or accidentally. The **jamming margin** is defined as

$$\text{jamming margin} = \text{process gain} - \text{signal-to-noise at output} - \text{implementation loss} \qquad \text{[Eq. 1]}$$

with the various terms expressed on a dB scale. This parameter determines the operating condition. A spread spectrum system cannot function if the level of interference exceeds the wanted signal by more than the jamming margin. It is impossible to design a system that would be immune from jamming. There is always a threshold and a spread spectrum process must be operated within this limit. However, there is a level of secrecy, without the need of encryption techniques, that cannot be provided by narrowband systems. This effect of signal concealment is one of the principal advantages of spread spectrum modulation.

The above parameters do not appear in the specification, but they would be considered at the design stage of the device.

(3) Radiofrequency technical characteristics

WWDS is a low power, wideband radio system intended for operation over short dis-

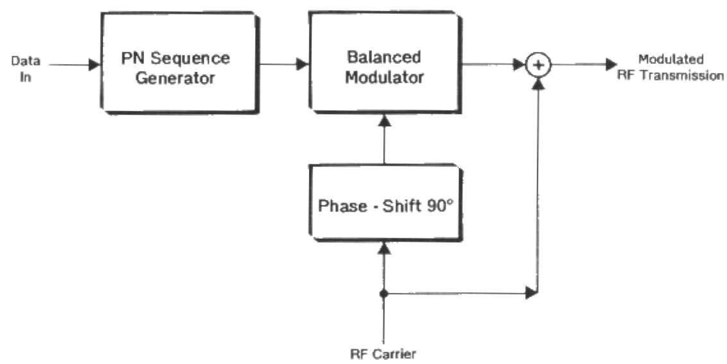


Fig. 1. Direct sequence spread spectrum modulation using phase shift keying (BPSK).

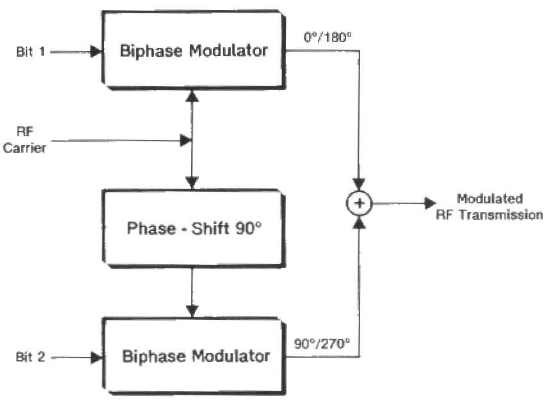


Fig. 2. Quadriphase modulation process.

tances. The aggregate transmission rate is 250 kbits per second. Units can be fixed (that is, base), mobile or hand portable. A mobile unit would be for installation in a vehicle. An antenna can be fitted as an integral part of a unit. Alternatively, it may have an antenna socket (50 Ω) to permit connection to an external antenna. The specification allows for three main categories of unit.

(a) Devices such as personal computers or data processing equipment that includes an optional printed circuit board(s) for WWDS operation. In this case, a WWDS unit is fitted internally within a host unit.
(b) Radio equipment that includes a facility for WWDS operation.
(c) WWDS units as separate devices.

Spread spectrum modulation is a requirement. If a manufacturer opts to use a modulation method other than Direct Sequence or Frequency Hopping, the technical requirements for Direct Sequence will apply. These points are discussed in greater detail in Sections 4 and 5. The frequency band is 2.4–2.5 GHz, but the operating frequency range is determined by the two frequencies where the **Power Spectral Density** (Appendix 3) is at −80 dBm Hz^{-1}.

The specification has three transmitter tests as follows. The methods for conducted measurements, using direct connections to test instruments, are explained. However, the reader should refer to the actual specification for the requirements for radiation measurements. For type testing, a manufacturer must provide a data sequence which spreads the output power through the power envelope. In the case of frequency hoppers, a manufacturer must supply a means of selecting the lowest and highest frequencies.

(i) **Effective radiated power**. The maximum value should not exceed 100 mW or 20 dBm. The term dBm refers to milliwatts measured on a dB scale referenced to 1 mW or $10\log_{10}$(ERP in milliwatts). However, it is not the practice to include W after the m. The test is carried out with modulation applied. The measurement method uses a diode detector and an oscilloscope of suitable bandwidth. The ERP is calculated from this result and the antenna gain.

(ii) **Power spectral density**. The maximum value should be 20 dBm in 100 kHz for Frequency Hopping and 10 dBm in 1 MHz for Direct Sequence or any other form of spread spectrum modulation. The test is performed by connection to a spectrum analyser or power measuring receiver through a peak hold detector. The test instrument is centred on the maximum value which is measured in dBm Hz^{-1}. Then the value for a 100 kHz or 1 MHz bandwidth is calculated. There is a requirement that a transmitter must be operating for at least 10 microseconds before the measurement is taken. If this is not the case, the test method must be included in the report for type approval of a unit.

(iii) **Spurious emissions**. The spectrum outside the operating range is monitored for emissions that could cause harmful interference to other services. The test is carried out with the modulation on. In the case of Frequency Hopping equipment the hopping is turned off. The specification gives a table of limits for different bands. If a unit has a stand-by mode, the test is repeated for this mode.

For a receiver there is only one test: Spurious Emissions as in (iii) but in a bandwidth of 120 kHz. Most radio receivers for test purposes have an IF bandwidth of 120 kHz and this test should be straightforward. However, a unit must be in the receive mode and there are different limits.

Tests for electromagnetic compatibility as specified by the EMC Directive 89/336 are not covered in the specification. In order to have the CE Trade Mark, a unit must be type approved to a separate standard for EMC tests.

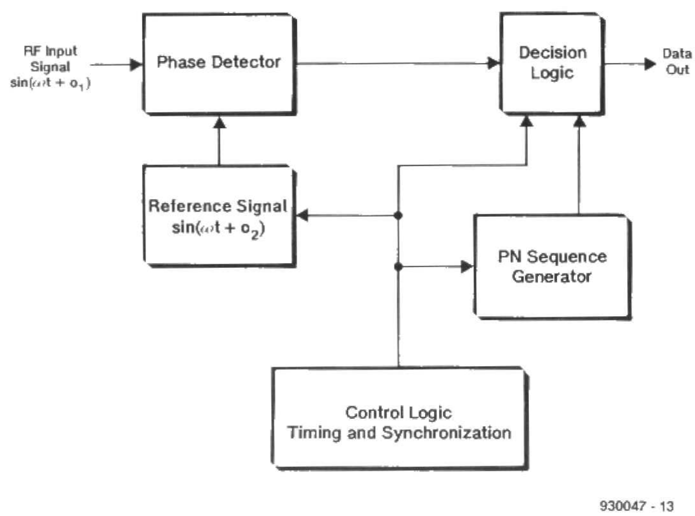


Fig. 3. Receiver system for direct sequence spread spectrum modulation.

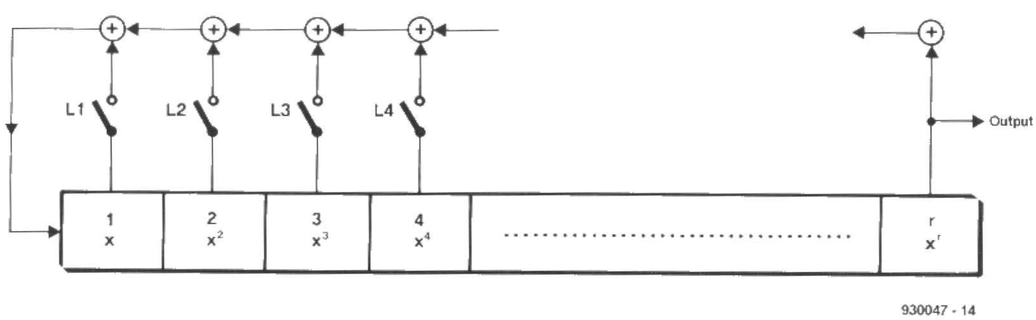


Fig. 4. Linear feedback shift register of *r* stages. The latch for the last bistable (flip-flop) is always closed in a maximum length arrangement. The term *x* represents a delay for successive stages.

(4) Direct Sequence Spread Spectrum Modulation.

The method is illustrated in Fig. 1 and PSK is the modulation process. The PN sequence generator outputs a sequence of bits such that the carrier undergoes a phase shift of 180° for a '1' and 0° for a '0'. The original data is not the modulating signal for the transmitter as explained in Section 2. Each data bit is combined with the PN sequence such that the output sequence constitutes a baseband signal for the modulation process. The technique of phase shifting is produced in two major steps. The balanced modulator mixes the two inputs to produce a product of these two signals. The same carrier shifted by 90° is added to generate the modulated output. Figure 2 shows an arrangement for two bits. The top circuit uses phase shifts of 0° and 180°, while the bottom uses 90° and 270°. This doubles the rate bit because each radiofrequency waveform represents two bits.

Figure 3 shows the process at the receiver. The code generator inputs the PN sequence without the data to the Decision Logic part of the circuit. The baseband signal is reproduced by the Phase Detector. An ideal analogue multiplier behaves as a phase detector. The output con-

tains a term $\sin(\theta_1-\theta_2)$. If the reference signal is applied correctly with $\theta_2=0$, this leaves θ_1 which is the original phase shift. The diagram is a simplified explanation as the shifts must be converted back to bit form for processing by the Decision Logic. Obviously, synchronization with a transmitter is very important as the carrier is effectively reinserted. Further development of these requirements is outside the scope of this paper.

The term PN means pseudo-noise and refers to the fact that the sequence appears random. The usual method of generation is with linear feedback shift registers. Figure 4 shows a circuit for a shift register of five stages. Each stage is a JK bistable (flip-flop). The initial state of the shift register is known as the seed. Each state generates another state and eventually the seed is produced again. Consequently, there is a specific period that is determined by the arrangement of the latches. The **Characteristic Equation** for r stages is

$$f(x)=1+\sum_{i=1}^{r}L_i x^i \qquad \text{[Eq. 2]}$$

and $f(x)$ is a factor of (x^p+1), where p is the period of the shift register. For a maximum length, $p=(2^r-1)$ and the output PN sequence has the same period. The arrangements that result in maximum length are used to generated PN sequeces. The properties of such sequences are as follows.

(a) Every state of the shift register (other than all zeros which is not used for obvious reasons) occurs once during a period.
(b) Every state has a unique predecessor and successor.
(c) The PN sequence has exactly 2^{r-1} '1s' and $2^{r-1}-1$ '0s'.
(d) The autocorrelation function is two valued with an in-phase and an out-of-phase value.

There are $\phi(2^r-1)/r$ arrangements for maximum length, where ϕ is **Euler's Totient Function.** For $r=5$, $\phi(31)/5=6$. The factors are

$$(x^{31}+1)=(x+1)(x^5+x^2+1)(x^5+x^3+1)(x^5+x^3+x^2+x+1)(x^5+x^4+x^2+x+1)(x^5+x^4+x^3+x+1)(x^5+x^4+x^3+x^2+1).$$

The example in Fig. 4 has (x^5+x^2+1) as it Characteristic Equation. Each factor is irreducible in that it cannot be factored further over GF(2) while keeping the coefficient of every power of x either '1' or '0'. The incoming bit on the left side is generated according to the equation

$$a_n=(a_{n-2}+a_{n-5}) \bmod 2. \qquad \text{[Eq. 3]}$$

It takes five shifts for this bit to pass through the register as per the conditions. Appendix 4 lists the sequence of states starting with the state (0,1,0,1,0). The output sequence would be the bits going down on the right side of the table. As per the above properties, there are 15 '0s' and 16 '1s'. A different seed would result in a time-shifted version of the same output sequence and of the sequence of states. A different characteristic equation (e.g., x^5+x^3+1) would still generate a maximum length sequence, but the states would occur in a different order. The reader is referred to Ref. 2 for a detailed analysis of shift register sequences.

From the point of the radio frequency spectrum, Direct Sequence using PSK digital modulation results in a power envelope of the form $[\sin(x)/x]^2$. Although the data specifically generates phase shifts of the radio frequency carrier, the process still produces sidelobes. The first sidelobe is 13 dB down on the main transmission. This is a waste of energy in that sidelobes are not necessary for the transfer of information. In addition, they can cause interference to other services. Consequently, to minimize the effects of sidelobes, other forms of modulation have been developed. In **Minimum Shift Keying** (MSK), the first sideband is 23 dB down, which is a considerable improvement. GMSK is used in cellular telephone and in DSRR. However, these points need not be considered in this type of overview.

(5) Frequency Hopping Spread Spectrum Modulation

In this method, illustrated in Fig. 6, a pseudo-random sequence is used to control a frequency synthesizer. The code generator is in fact a PN Sequence Generator as in Fig. 4. During operation, the synthesizer generates different frequencies such that the frequency of the transmission is not constant but hops through a specific set of values over a large frequency range. The code generator does not produce a baseband signal for the modulator as in Fig. 1, but controls the hopping mechanism. At individual frequencies, the modulation process will produce sidebands; spurious emissions, in addition to the wanted signal, may also exist. The transmission at each frequency is the equivalent of a short rectangular pulse with a spectrum of the form $\sin(x)/x$ centred at that frequency. From the point of the radiofrequency spectrum, both methods of spread spectrum modulation are the same.

A receiver must generate the exact same sequence in order that its own synthesizer hops through the same frequencies. In addition, it must be synchronized with the transmitter to ensure that the hops occur at the correct time instants. Figure 7 shows a typical arrangement for a process using an intermediate frequency. The frequencies generated in the receiver are offset such that the difference between the two frequencies (i.e., the intermediate frequency) is constant. If f_j is the input frequency at the front end, the synthesizer produces (f_j+f_{if}) such that f_{if} is input to the demodulator. Both diagrams are deliberately simplified for the presentation of an overview. The reader is referred to Ref. 1 for a detailed analysis.

(6) Summary

The specification only covers the radio aspects of a system. Coding requirements, such as error correction or data compression techniques, are left to the manufacturer. It could be argued that there is no real requirement in these areas as a WWDS devices would be working over short distances on low power. The specification covers the minimum standard for operation while minimizing any unwanted effects to other services. The above comments are purely personal and a reader should consult the specification for the exact position.

References

1. *Spread Spectrum Modulation*, by Robert C. Dixon, John Wiley & Sons (1984)

2. *Shift Register Sequences*, by Solomon W. Golomb, Aegean Park Press (1982)

3. *Telecommunications Engineering*, by H.G. Brierly, Edward Arnold (1986)

Appendix 1 Data Transmission

The main technical requirements are:

1. Transmission rate of 1200 bits per second using fast Frequency Shift Keying (FSK).
2. Modulation rate of 1200 baud. This is the signalling rate in terms of the number of electronic messages per second. For example, telex machines operate at 50 baud, which means that every bit has a time slot of 20 milliseconds. A charac-

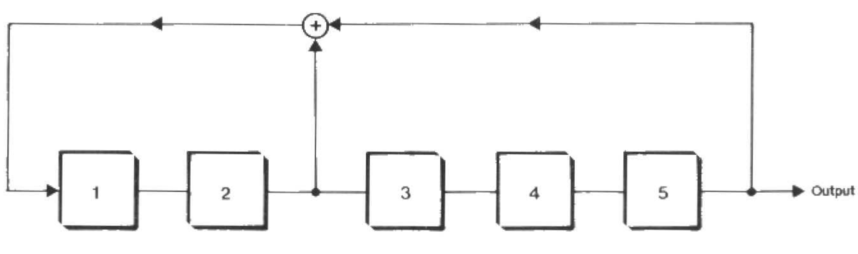


Fig. 5. Linear feedback shift register to generate a PN sequence of period 31.

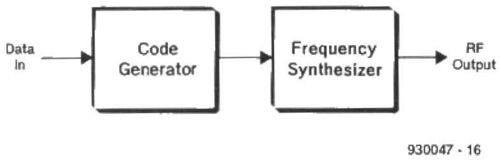


Fig. 6. Frequency hopping spread spectrum modulation.

ter is represented by 5 bits plus 1.5 start bits and 1 stop bit. Since every signal is represented by one bit in 2-tone FSK, the two rates of transmission and modulation are equal.
3. A '0' is represented as an 1800 Hz tone.
4. A '1' is represented as a 1200 Hz tone.
5. Every bit starts with a phase shift of 0° or 180°.

Appendix 2
Power Flux Density

This is the power per unit surface area and the units are watts per square metre (W m^{-2}). If the output power from a transmitter is P watts, the PFD at a distance of d metres is $PFD=P/(4\pi d^2)$. If the antenna gain of the transmitter is G, P is replaced by PG in the above equation. The term PG is known as the **Equivalent Isotropic Radiated Power (EIRP)**. It is common to use **Effective Radiated Power (ERP)**, where G is referenced to a tuned dipole. (The difference is 1.5 dB, which is the gain of a dipole referenced to an isotropic antenna.) It is usual to express the PFD in dBW m^{-2}, which means that the above value becomes $10\log_{10}$ of that value.

$$PFD \text{ in dBW m}^{-2} = PG \text{ in dBW} - 10\log_{10}(4\pi) - 20\log_{10}(d).$$

For example, PFD limits are used in the Radio Regulations to calculate levels of interference between different services. Article 28 gives specific limits to minimize the possibility of harmful interference from satellites to terrestrial services above 1 GHz. RR2557 gives a maximum value of –144 dBW m^{-2} in any 4 kHz bandwidth for arrival angles between 25° and 90°.

The **Electric Field Strength** is given by

$$PFD=E^2/Z_0,$$

where $Z_0 = 377\ \Omega$ is the **Impedance of Free Space**.

$$20\log_{10}E = 10\log_{10}PFD + 10\log_{10}377,$$

$$\therefore E \text{ in dBV m}^{-1} = PFD \text{ in dBW m}^{-2} + 10\log_{10}377.$$

Since E is usually expressed in microvolts per metre (μV m^{-1}) in radio communications, the above equation can be rewritten as

$$E \text{ in dB}\mu\text{V m}^{-1} = PFD \text{ in dBW m}^{-2} + 10\log_{10}377 + 120.$$

Thus –80 dBW m^{-2} would be equivalent to 66 dBμV m^{-1}. A simple conversion is to add 146 to the PFD in dBW m^{-2}. If the distance is expressed in kilometres (km) or the power in kilowatts (kW), the equations must be adjusted accordingly.

Appendix 3
Power Spectral Density

This is the power in terms of frequency rather than area as per Appendix 2. The units are watts per Hertz (W Hz^{-1}). Thus, in simple terms, if the Power Spectral Density is PSD W Hz^{-1} between two frequencies, f_1 and f_2, the power is given by

$$P = PSD(f_2 - f_1).$$

In reality, the PSD would be a function of frequency, that is, $P_s(f)$, and a more complicated mathematical operation would be required as follows

$$P = \int_{-\infty}^{\infty} P_s(f)\,df$$

If $P_s(f)$ is zero outside the range f_1 to f_2, the limits of integration would be adjusted from $-\infty$ to f_1 and from $+\infty$ to f_2. An actual expression for $P_s(f)$ within the operating band could be quite complicated and the integration operation could be difficult.

In the article, PSD is used rather than PFD to specify the technical requirements at radio frequencies. Obviously, this is more appropriate for spread spectrum modulation where emissions are deliberately spread through a wide frequency band and are not confined to narrow channels.

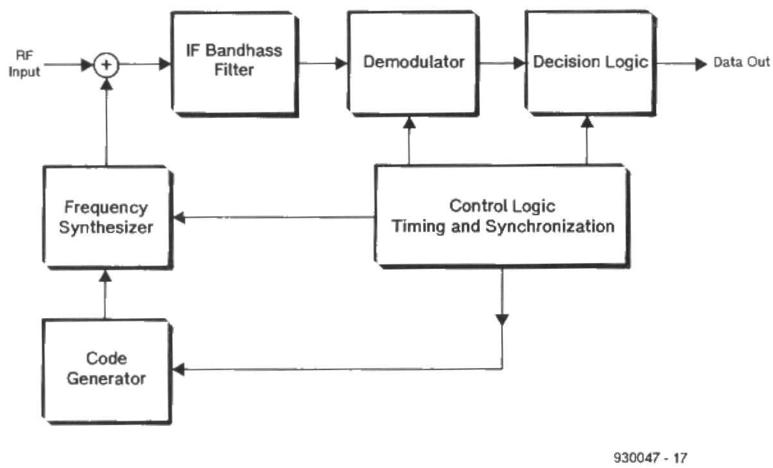


Fig. 7. Receiver process for frequency hopping spread spectrum modulation.

Appendix 4
PN Sequence Generator

For the arrangement of Fig. 4, the cycle of states for a seed or initial state at stage 1 is as follows.

Number	State of Shift Register	Output
1	0 1 0 1 0	0
2	1 0 1 0 1	1
3	1 1 0 1 0	0
4	1 1 1 0 1	1
5	0 1 1 1 0	0
6	1 0 1 1 1	1
7	1 1 0 1 1	1
8	0 1 1 0 1	1
9	0 0 1 1 0	0
10	0 0 0 1 1	1
11	1 0 0 0 1	1
12	1 1 0 0 0	0
13	1 1 1 0 0	0
14	1 1 1 1 0	0
15	1 1 1 1 1	1
16	0 1 1 1 1	1
17	0 0 1 1 1	1
18	1 0 0 1 1	1
19	1 1 0 0 1	1
20	0 1 1 0 0	0
21	1 0 1 1 0	0
22	0 1 0 1 1	1
23	0 0 1 0 1	1
24	1 0 0 1 0	0
25	0 1 0 0 1	1
26	0 0 1 0 0	0
27	0 0 0 1 0	0
28	0 0 0 0 1	1
29	1 0 0 0 0	0
30	0 1 0 0 0	0
31	1 0 1 0 0	0
32	0 1 0 1 0	0

After 31 stages the sequence is reproduced again. Consequently, the period is $31=2^5-1$ and is a maximum length sequence as predicted.

An interesting point is that the last latch is closed for all six factors. This is not a coincidence but is due to the fact that the auto correlation function is two-valued. This holds for the shift registers of any length. ■

READERS' CORNER

LETTERS

Dear Editor—I am interested in making the Horn Loudspeaker (*Elektor Electronics*, May 1990), but I feel that several questions need to be answered before I could begin.

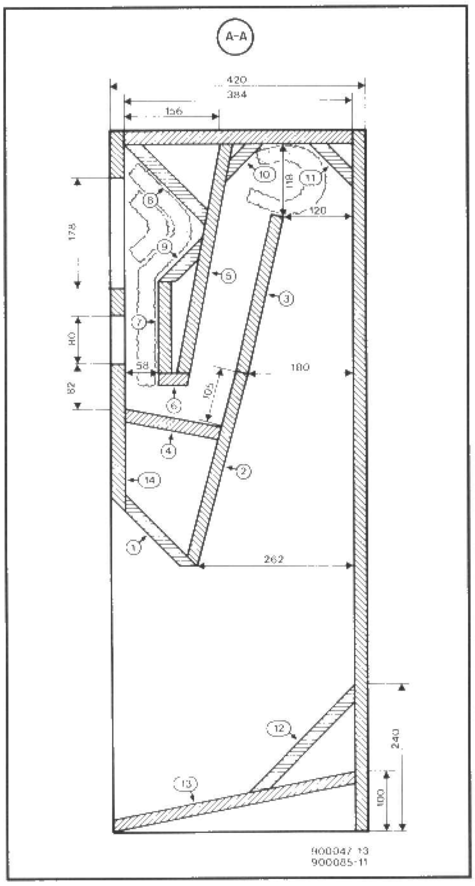
1. When asking suppliers for McFarlow speaker units, their answer is 'Never heard of them'.
2. More information is required regarding the dimensions of the horn to give the possibility of modifying to take different drive units. Also the compromises and aims of the designer.
3. The drawing of Fig. 7 shows the total width as 250 mm and the internal as 214 mm, whereas the cutting list shows these as 288 mm and 250 mm respectively. Which is correct?
4. No mention is made of the finished product's sound quality, e.g., comparison to other types of speaker. Although test graphs are supplied, more information would be appreciated.
5. As regards 'the use of expanded polystyrene in the speaker with instructions to "fill the enclosure"', I understand that a horn should be empty for the free passage of the sound.

K. Halsall, Warrington.

Your letter is typical of a number we have received over the past few years. Since interest in this project remains high, a number of points that have been published, or have been included in individual answers, are repeated here.

1. The McFarlow drive units are fairly well known—as reported in various past issues (the first in June, 1990), these units are available from Bochako, 46 London Road, Kingston-upon-Thames KT2 6QF; Telephone 081 541 4433; Fax 081 547 1096.

2. The designer replies that his aim was a design based entirely on the McFarlow units stated in the article. There were no compromises nor comparisons.



3. Corrections on several dimensions in Fig. 7 were published before as follows: total width of front panel is 286 mm; internal width is 250 mm; in view A-A, the dimension of 129 mm given at the top should be 156 mm; in the same view, the dimension of 240 mm at the right-hand bottom pertains to the top, not the lower, edge of panel 12.

4. See 2.

5. The placing of damping material was unfortunately not shown in Fig. 7 (but is shown in the drawing here); moreover, the instruction should have read: 'Three pieces of wad-ding, (a) 20×30 cm; (b) 25×30 cm; (c) 12×20 cm (width×length), should be placed as shown in the figure. Piece (a) is folded along its length and clamped between panels 3, 10 and 11. Piece (b) is laid against panels7, 8 and 9. It need not be glued into place, because the woofer will hold it in place. Piece (c) is put into place through the woofer opening against piece (b). This is best done just before the woofer is fitted (lay the cabinet on its back). Press the woofer well into the wadding, otherwise it may prove difficult to screw it into place.

[Editor]

Dear Editor—I made a modification to the excellent 'Video Digitizer' (July/August 1991) that other readers may be interested in. It consists of fitting an on-board 24 MHz crystal oscillator to lose that awkward flying lead. If the optional I/O port is not fitted, the oscillator module can be fitted in place of IC_{15}. Splay the supply leads to fit into pin 8 and 16 and bend the output pin of the module under the body so that a small wire can be soldered on and connected to terminal 'T' on the dig-

EVENTS

IEE AND IEEIE PROGRAMME

29 Mar1 Apr—Developments in power systems protection.
30 Mar–2 Apr—Antennas & propagation.
1 Apr—10th Lord Nelson of Stafford Lecture.
1 Apr—Electromagnetic compatibility for project engineers.
2 Apr—Advances in neural networks for control and systems.
4–7 Apr—Safety critical systems.
6 Apr—Very high speed IC hardware description language.
6 Apr—Aspects of telecommuting.
7 Apr—Total quality management.
14 Apr—Quality assurance (and QC).
14–16 Apr—UK teletraffic.
15 Apr—Motors and drives for battery powered propulsion.
16 Apr—Optical detectors and receivers.
18–21 Apr—Telecommunications.
18–22 Apr—Digital techniques in radio systems.
19 Apr—Engineer your own success.
19 Apr—Design calculations for small electrical installations.
19 Apr—Laser applications.
19–22 Apr—Spectroscopy: the changing face of physics.
19–24 Apr—Test.
20 Apr—The engineer's role in customer care.
21 Apr—Improved sound for high definition television.
21 Apr—Performance measures.
21 Apr—Professional negligence of engineers.
21 Apr—Education and training across frontiers.
21–23 Apr—Multimedia technologies and future applications.
22 Apr—Case studies in EMC.
22 Apr—Home electronic systems.
22 Apr—Recent developments in optoelectronic materials and devices.
22–23 Apr—Grammatical inference.
27 Apr—Spread spectrum techniques for radio communication systems.
27 Apr—Portable appliance testing.
28 Apr—Watson-Watt lecture.
28 Apr—Brakes and braking systems.
29 Apr—Analogue computation and simulation: a history.

Further information on these, and many other, events may be obtained from the IEE, Savoy Place, London WC2R 0BL; Telephone 071 240 1871, or from the IEEIE, Savoy Hill House, Savoy Hill, London WC2R 0BS; Telephone 071 836 3357.

All Formats Computer Fairs will be held on 3 April in Edinburgh, Appleton Tower, George Square; 4 April, Glasgow, City Hall; 17 April, Nottingham, Jesse Boot Centre, University; 18 April, West Midlands, National Motorcycle Museum, J6 M42; 24 April London, Sandown Park, Esher, J9/10 M25; 25 April, Bristol, Brunel Centre, Templemeads. Further information from Bruce Everiss on (0608) 662 212.

The **Cable & Satellite Exhibition 93** will be held in London on 5–7 April. Further information from Reed Exhibitions on 021 705 6707.

itizer board. Pin 1 of the oscillator is not connected internally, and can be fitted into pin 1 on the PCB to provide mechanical stability. If the I/O port is fitted, simply solder the module to the pins of IC_{15} once in situ. This mod is particularly useful for owners of the A3000 who have to use the rear expansion port

It seems that most computer-based projects are geared to the PC market, so it was encouraging to see one for the Archimes range of computers. I had hoped that Mr Kortink's article would spark off a speight of Arc-based projects, but so far I have seen nothing.

What I would like to see is an EPROM programmer for the Arc as I have contacted various sources and there does not seem to be one available at all. Perhaps the 8052-BASIC computer featured in *EE* in 1992 would do the job admirably since it has an on-board EPROM programmer.

J.S. Patterson, Newcastle-upon-Tyne.

Unfortunately the response to the Video Digitizer project has been much smaller than expected; apparently there are not too many Archimedes users around who are interested in adding hardware to their machine (this also seems to apply to Amiga users). We hope this letter may change that.

Your suggestions regarding an EPROM programmer have been passed to our Central Design Department. [Editor]

Figuring it out

The solution to the exercise on p. 29 is: Taking point E as the reference point for zero volts, the potentials at other points are: A = 5 V; B = $3^1/_3$ V; C = 2 V; D = 4 V.

COMPONENT RATINGS

In resistor and capacitor values, decimal points and large numbers of zeros are avoided wherever possible. Small and large values are usually abbreviated as follows:

p (pico-) $= 10^{-12}$
n (nano-) $= 10^{-9}$
µ (micro-) $= 10^{-6}$
m (milli-) $= 10^{-3}$
k (kilo-) $= 10^{3}$
M (mega-) $= 10^{6}$
G (giga-) $= 10^{9}$

Note that nano-farad (nF) is the international way of writing 1000 pF or 0.001 µF.
Resistors are $^1/_3$ watt, 5% metal film types unless otherwise specified.

The direct working voltage of capacitors (other than electrolytic or tantalum types) is assumed to be ≥60 V. As a rule of thumb, a safe value is about 2× direct supply voltage.

Direct test voltages are measured with a 20 kΩ/V meter unless otherwise specified.

Mains (power line) voltages are not listed in the articles. It is assumed that our readers know what voltage is standard in their part of the world.

Readers in countries that use 60 Hz supplies, should note that our circuits are usually designed for 50 Hz. This will not normally cause problems, although if the mains frequency is used for synchronization, some modification may be required.

The international letter symbol '*U*' is used for voltage instead of the ambiguous '*V*'. The letter V is reserved for 'volts'.

The size of a metric bolt or screw is defined by the letter M followed by a number corresponding to the overall diameter of the thread in mm, the × sign and the length of the bolt or screw, also in mm. For instance, an M4×6 bolt has a thread diameter of 4 mm and a length of 6 mm. The overall diameter of the thread in the BA sizes is: 0 BA = 6.12 mm; 2 BA = 4.78 mm; 4 BA = 3.68 mm; 6 BA = 2.85 mm; 8 BA = 2.25 mm.

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INFRA-RED RECEIVER FOR 80C32 SINGLE-BOARD COMPUTER

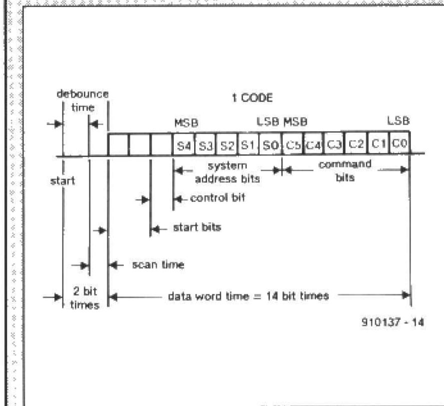
This article describes a sensitive infra-red receiver that enables an 80C32-based microcontroller system to interpret commands sent by an RC5 compatible remote control transmitter. Test software is available!

Design by W. Hackländer

RC5-COMPATIBLE infra-red remote control transmitter units are widely available as spare parts for TVs, video recorders and hi-fi equipment. Here, we propose an RC5 compatible receiver interface for our 80C32 single board computer (SBC) plus extension card (Ref. 1). Although the present receiver is specifically designed to link to this microcontroller board, it has a databus buffer that makes connection to other systems rather simple. It is assumed here that you are familiar with the operation of the RC5 infra-red remote control system (Fig. 1). If not, you will find extensive background information in Ref. 2.

How it works

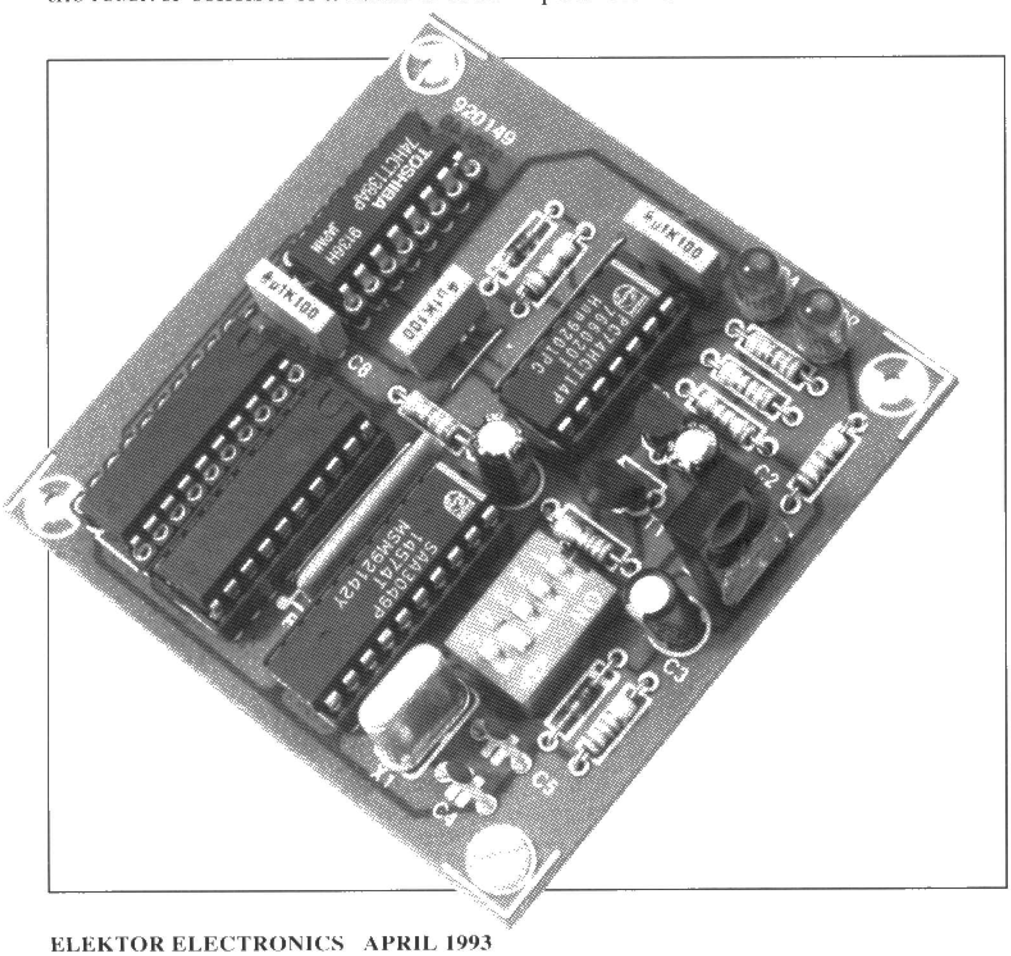
The circuit diagram (Fig. 2) shows that the receiver consists of a handful of integrated circuits only. The infra-red signal is received and amplified by IC1. The decoder, IC2, first checks if the signal contains valid RC5 codes. If so, the received address is compared with the address set by switch S1. If the addresses match, i.e., a code is received that is intended for this receiver, the decoder puts the data on the six data outputs, and supplies a short pulse on pin 19. This pulse arrives at LED D2 via two transistor stages. The LED lights to acknowledge reception. The pulse is also 'stretched' via inverter IC3a and network D3-R8-C6. This is done to 'bridge' the time between two consecutive pulses, if the associated key on the transmitter is kept pressed for some time (e.g., for volume control). The 'stretched' pulse is buffered by an inverter (IC3b) before it arrives at inputs A6 and A7 of bus buffer IC4.



System address	Equipment
0	TV
2	Teletext
5	video recorder
7	experimental
16	preamplifier
17	receiver/tuner
18	tape/cassette recorder
19	experimental
20	CD player

Command number	Command
0-9	0-9
12	standby
13	mute
14	preferences
16	volume +
17	volume –
18	brightness +
19	brightness –
20	colour saturation +
21	colour saturation –
22	bass +
23	bass –
24	treble +
25	treble –
26	balance right
27	balance left
48	pause
50	fast reverse
52	fast forward
53	play
54	stop
55	record

Fig. 1. Overview of RC5 command numbers and addresses. The desired receiver address is set with the aid of switch S1. The controller reads the command number as the six least significant bits of IC4.



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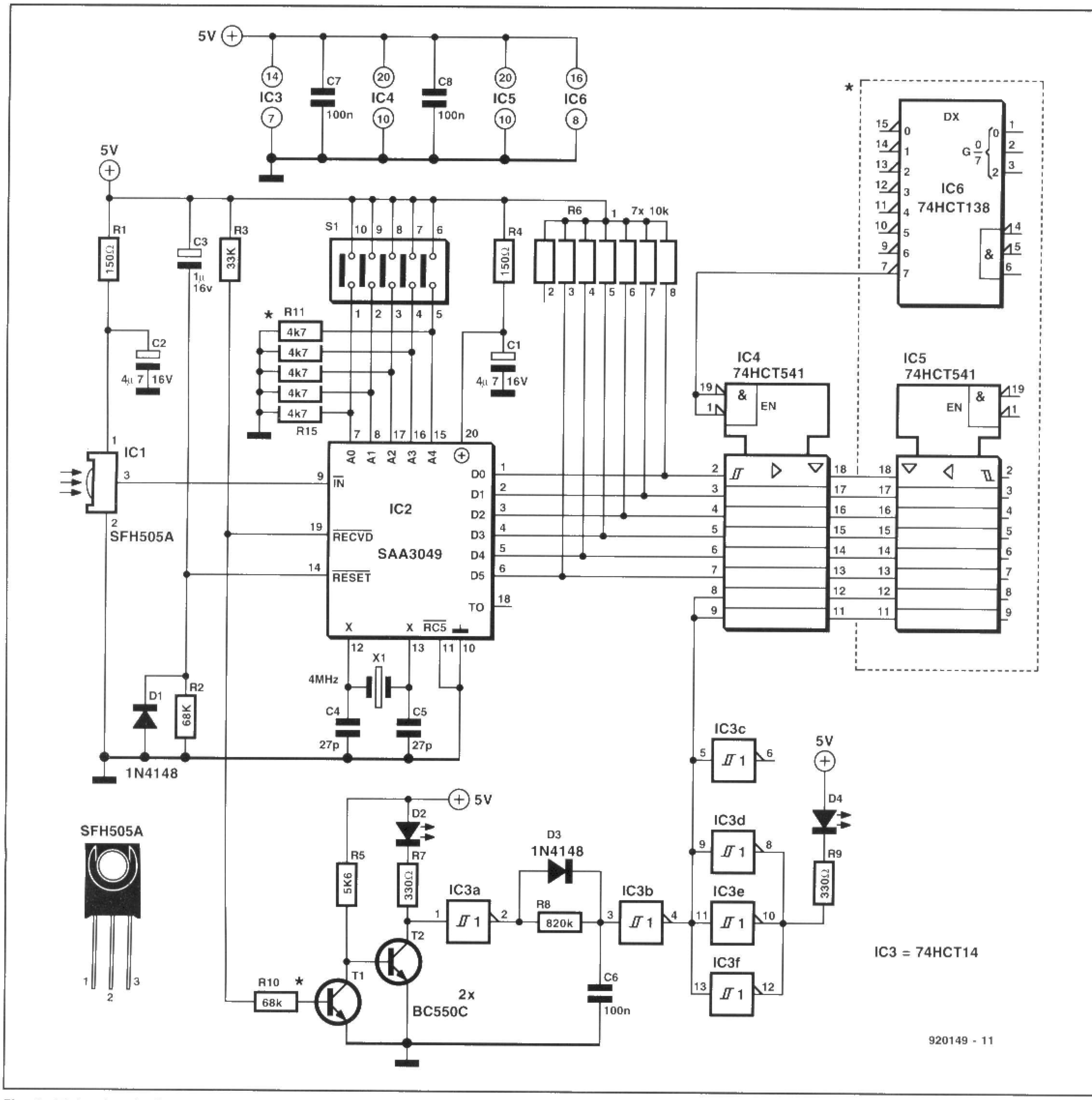

Fig. 2. This circuit diagram shows that a sensitive infra-red receiver can be built from a handful of components.

These two bits enable the microcontroller to detect that the other six bits contain fresh data, and how long that data is being transmitted (0 = data received; 1 = no data received). In addition, the output signal of IC3b is also indicated visually by an LED (D4). The drive current for the LED is supplied by the three remaining inverters in IC3.

Circuits IC5 and IC6 are actually part of the 80C32 extension card, although they are included in the circuit diagram of the infra-red receiver. IC4 is connected to the data bus on the extension card via the pins of IC5 (= IC3 on the extension card). The chip select signal for IC4 is supplied by address decoder IC6 (= IC2 on the extension card). The IC sockets marked IC5 and IC6 on the infra-red receiver board have extra long pins, which are inserted into the sockets for IC2 and IC3 on the 80C32 extension card. The two integrated circuits, IC2 and IC3, originally fitted on the extension card, are simply moved to the infra-red receiver board. Evidently, if you have a different microcontroller system, it is equally simple to connect it to the receiver via pins 11-18 of IC5, and pin 7 of IC6.

Finally

The printed circuit board design is given in Fig. 3. Unfortunately, resistors R10-R15 are missing. They are, however, easily fitted at the solder (track) side of the board. Before fitting R10, cut the track that runs from R3 to the base of T1. R10 is then soldered across this interruption. Resistors R11-R15 are connected to switch S1 and ground as per the indications in the circuit diagram.

Be sure to use IC sockets with extra long pins (for instance, wire-wrap types) in positions IC5 and IC6. If necessary, the pins may be lengthened by stacking a couple of standard IC sockets.

The most frequently used addresses and commands contained in the RC5

COMPONENTS LIST

Resistors:

2	150Ω	R1;R4
2	68kΩ	R2;R10
1	33kΩ	R3
1	5kΩ6	R5
1	7-way 10kΩ SIL	R6
2	330Ω	R7;R9
1	820kΩ	R8
5	4kΩ7	R11-R15

Capacitors:

2	4μF7 16V radial	C1;C2
1	1μF 16V radial	C3
2	27pF	C4;C5
3	100nF	C6;C7;C8

Semiconductors:

2	1N4148	D1;D3
2	red LED	D2;D4
2	BC550C	T1;T2
1	SFH505A* (Siemens)	IC1
1	SAA3049** (Philips Semiconductors)	IC2
1	74HCT14	IC3
2	74HCT541	IC4;IC5
1	74HCT138	IC6

Miscellaneous:

1	5-way DIP switch block	S1
1	4 MHz quartz crystal	X1
1	Printed circuit board and software (1791); set code 920149 (see page 70).	

* Siemens distributor: ElectroValue, 28 St Jude's Road, Englefield Green, Egham, Surrey TW20 0HB. Telephone: (0784) 433603. Fax: (0784) 435216.

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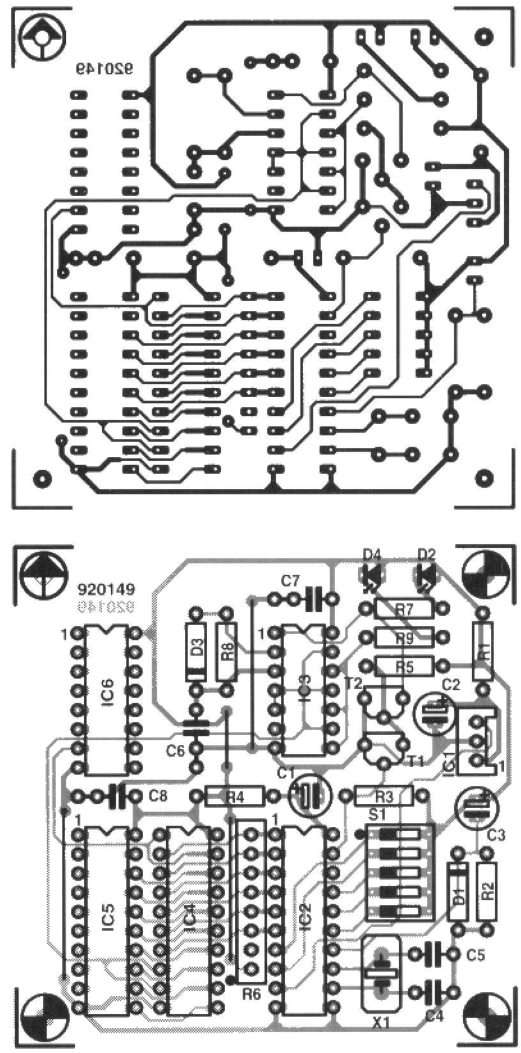
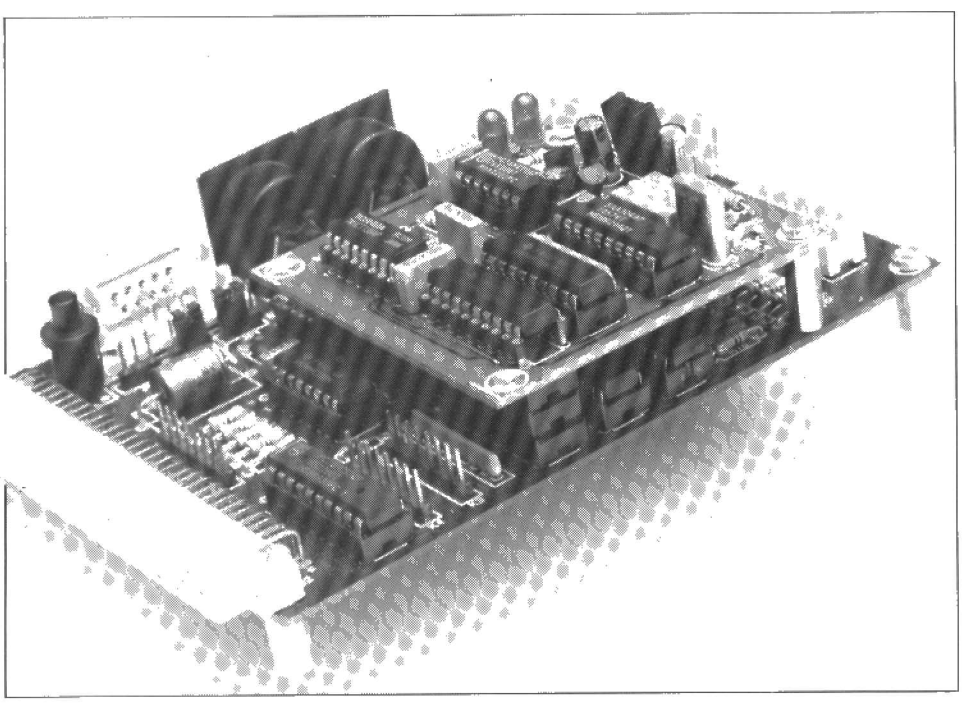


Fig. 3 The receiver board is designed to fit 'piggy back' on the 8032 SBC extension card.

set are listed in Fig 1. This information allows you to design an application that meets the RC5 standard as far as possible. It also tells you the codes that are best not used (to avoid interference on other RC5 compatible equipment).

As to software, a test routine is available for the combination (1) infra-ed receiver described here; (2) 80C32 single-board computer; (3) 80C32 extension card; and (3) an RC5-compatible remote control transmitter (for instance, a unit from the Policom series). This test routine is run from a floppy disk supplied through the Readers Services. Apart from testing the receiver, you may also use the MCS-51 source code on the disk as a starting point for further software development, or for your own experiments. The source code is written for the EASM-51 assembler, which is used for our MCS-51 assembly language course (Ref. 3). ■



References:

1. 8052/80C32 single-board computer. *Elektor Electronics* May 1991.
2. Universal RC5 code infra-red receiver. *Elektor Electronics* January 1992.
3. 8051/8032 Assembler course. Eight instalments, starting in *Elektor Electronics* February 1992.

Temperature-insensitive voltage divider

By Bryan Hart, BSc, CEng, MIEE

Introduction

In equipment design it is often necessary, for accurate voltage-level comparisons, to generate a temperature-insensitive voltage level U_R that is precisely related to an input voltage U_i or the rail supply U_{cc}. In particular, the convenient choice $U_R = U_{cc}/2$ leads to the standard formula $t_d = 0.69 \times$(time constant) for the time delay, t_d, in some types of monostable circuit. When it is necessary to supply significant current from U_R, a simple resistive potentiometer divider network is inadequate and active devices have to be employed in the division process.

Circuit description

Figure 1 shows a proposed precision voltage-divider network, with low small-signal output resistance, that is capable of operating down to $U_i = 2$ V. The design exploits the close matching, and the temperature tracking, of the respective parameters of similar devices fabricated in close proximity on the same chip. Transistor T_3 is included to equalize the emitter currents of T_1 and T_2 despite variation in input voltage U_i.

If the resistors are perfectly matched (that is, $R_1 = R_2 = R_3 = R$) and the transistors are assumed to have identical characteristics, $I_{b1} = I_{b2}$; $\delta_1 = \delta_2 = \delta_3 = \delta$; and $I_1 = I_2 = I_3 = I$.

Simple circuit analysis then gives

$$U_R = \{(U_i - 2\delta)/2\} + \delta = U_i/2. \qquad \text{[Eq. 1]}$$

Eq. 1 is valid if T_2 does not saturate. This requires

$$U_{cc} \geq \{(U_i/2) + \delta\},$$

where $\delta \approx 0.7$ V.

Matching is imperfect in practice, so Eq. 1 must be modified to

$$U_R = (U_i/2)(1+\varepsilon_T) = (U_i/2)[1+(\varepsilon_R+\varepsilon_\beta+\varepsilon_s)] \qquad \text{[Eq. 2]}$$

In this, ε_T is the total fractional variation in U_R. Its components, ε_R, ε_β, ε_s, result from, respectively, resistor tolerances; mismatch in the common-emitter direct current gain, β; and finite base-emitter offset voltage, $U_{os} = (\delta_1 - \delta_2)$.

With the usual assumption that $\beta >> 1$, a straightforward (though tedious) algebraic analysis gives:

$$\varepsilon_R = \pm(\Delta R/R)$$

= fractional resistor tolerance;

$$\varepsilon_\beta \approx \pm(\Delta\beta/2\beta^2$$
$$= (1/2\beta)(\text{fractional mismatch in } \beta);$$

$$\varepsilon_s = \pm(U_{os}/U_i).$$

Note that ε_β increases as U_i gets smaller because β falls off as I_c decreases.

The error terms are shown as additive in Eq. 2. This is a worst possible case. However, the errors are all uncorrelated, so a more realistic estimate of ε_T can be obtained by assuming they add in a root-mean-square sense. Thus,

$$\varepsilon_T \approx (\varepsilon_R^2+\varepsilon_\beta^2+\varepsilon_s^2)^{1/2}. \qquad \text{[Eq. 3]}$$

The low frequency incremental output resistance, r_o, is given by

$$r_o \approx (R/2\beta)+(r_e/2), \qquad \text{[Eq. 4]}$$

in which $r_e\,(\Omega) \approx 25/I_c\,(\text{mA})$,

and $I_c = (U_i - 2\delta)/2R$.

R needs to be 'low' for low values of r_o and ε_β when U_i is at its minimum, but 'high' for acceptable circuit dissipation when U_i is at its maximum. Hence, the choice of ohmic value for R is a non-critical compromise.

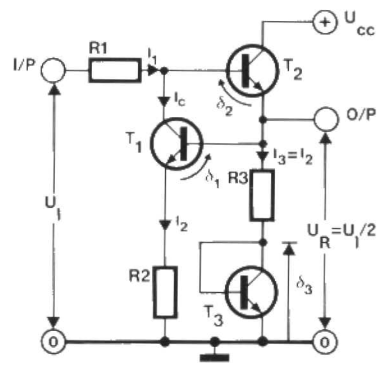


Fig. 1. Proposed precision divider network. Practical component choice: $R_1 = R_2 = R_3$ =3 kΩ ±1%, metal film; T_1, T_2, T_3= part XB101 (Exar).

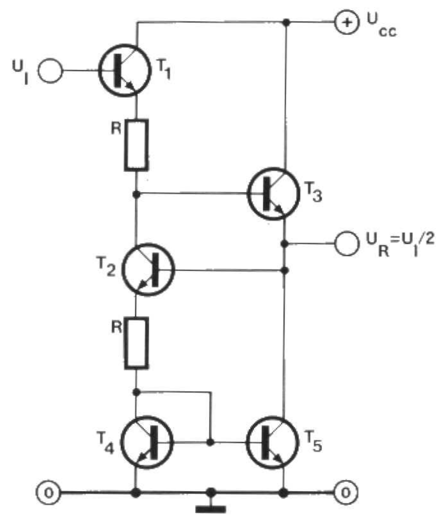


Fig. 2. Modified network with reduced source loading.

Circuit performance

The circuit in Fig. 1 was constructed from the components indicated. The Exar Kitchip part XB101 is a monolithic, small-signal n-p-n transistor array comprising five matched devices for which the specified data are:

$300 > \beta > 80$ at $I_c = 1$ mA;

$(\Delta\beta/\beta) \leq 0.1$; and

$U_{os}(\text{max}) = 6$ mV.

Since I_c falls below 1 mA in the proposed circuit, we will assume that $\beta(\text{min}) = 50$. Then, calculations give $\varepsilon_T(\%) = \pm 1$; $\varepsilon_\beta(\%) = \pm 0.1$; $\varepsilon_s(\%) = \pm 0.3$ at $U_i = 2$ V. Thus, in the worst case, $\varepsilon_T(\%) = \pm 1.4$: using Eq. 3, we obtain $\varepsilon_T(\%) = \pm 1.05$.

In practical measurements, U_i was connected to U_{cc} and varied over the range 2 V to 12 V. For 4 V$>U_i\geq$2 V, $\varepsilon_T(\%) < 0.35$; for 12 V$\geq U_i \geq$4 V, $\varepsilon_T(\%) < 0.2$.

With $U_i = 5$ V, U_R decreased by 15 mV for an external load of 0.45 mA.

In temperature tests, T_3 was used as an on-chip thermometer. The pd, δ_3, between its base and emitter terminals decreases by 60 mV when the surface of the DIL package was heated. Taking the usually quoted rate of change (≈ -2 mV °C^{-1}), this corresponds to a temperature rise of some 30 °C. This produced a change in U_R of less than 1 mV, corresponding to a temperature coefficient of output voltage of less than 15 p.p.m. over the measured range. This impressive performance results from

the close thermal coupling and base-emitter voltage tracking.

Similar performance is to be expected with other commercially available manufacturers' chips, e.g., the popular low-cost CA3046 transisto array and its equivalents. However, the best performance is to be expected from the likes of the MAT-04A/E matched-quad transistor package (Precision Monolithics) which is specified as having:

β>400 at I_c=10 µA; $(\Delta\beta/\beta)\leq 0.02$;

U_{0S}<0.2 mV.

As the circuit stands, T_2 could be destroyed by an accidental short-circuit to earth at the output. Such damage can be avoided by incorporating a small dissipation-limiting resistor, R_x, in the collector circuit of T_2. No figure is specified on the XB-101 data sheet for the maximum permitted power dissipation of an individual device, but if we assume a conservative value of 100 mW, R_x=560 Ω is suitable for $U_{cc}\leq 15$ V.

Making $R\leq 3R_x$, guarantees that T_3 operates in the linear-active mode, so that R_x plays no significant part in normal circuit operation, with an external load current up to twice the emitter bias current of T_1 and T_2.

Figure 2 shows a circuit variation that is appropriate when the loading of a source by the circuit has to be minimized. T_1 functions as an emitter-follower, while T_4 has two roles. It compensates for the base-emitter drop of T_1 and acts as the input stage of a simple 1:1 current mirror, the output of which is the collector current of T_5. Thus, the emitter current of T_3 is made sensibly equal to that of T_2 without the requirement of a third resistor. ■

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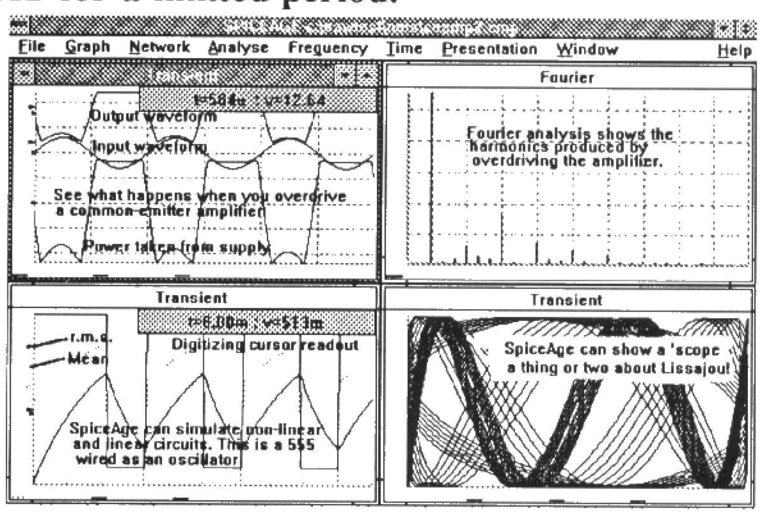

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4-MBYTE PRINTER BUFFER INSERTION CARD (PART 2)

Design by R. Degen and J. Dieters

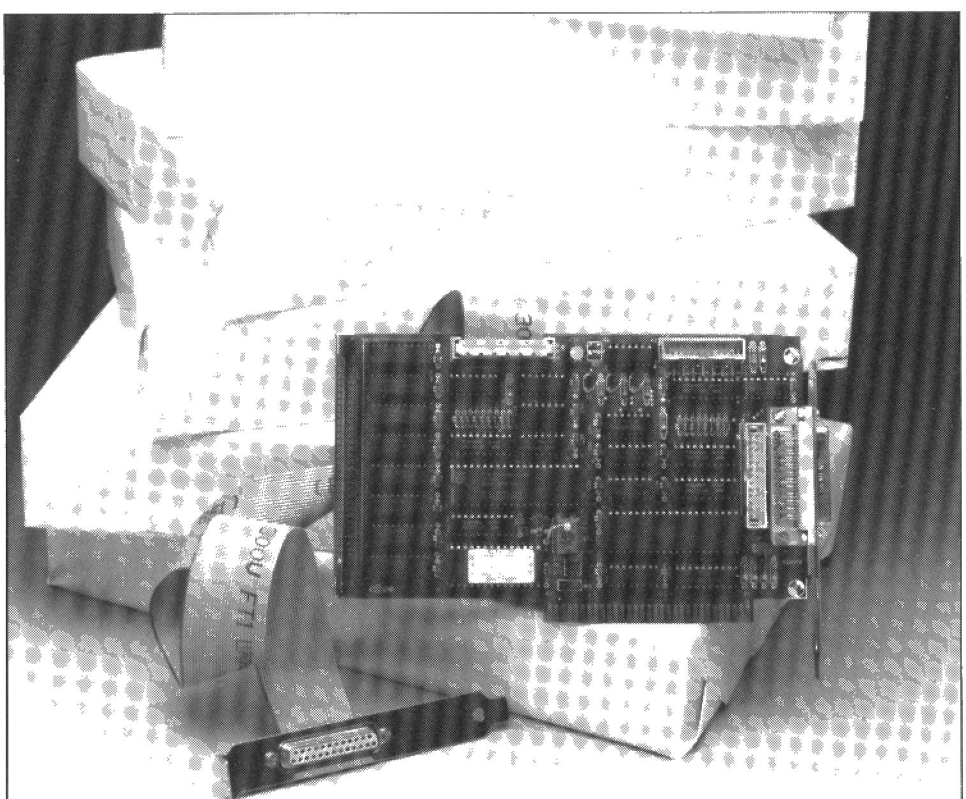
Construction

The printed circuit board designed for the printer buffer is a double-sided, through-plated type (Fig. 4) with gold-plated bus contact fingers. The track layouts are not given because of lack of space. We do not recommend making this PCB yourself, as you will easily run into trouble with the very high track density.

Start the construction by separating the main board and the display board. Let us concentrate on the display board first — few remarks are in order. The display board has on it three different types of 'Digitast' push-button: S1 is a switch with one make contact; S2 is a switch with one make contact and a built-in LED; and S3 is a switch with one make contact, locking action and two built-in LEDs. Make sure to fit the LEDs at such a height that the push-buttons are easily operated. Pin header K3 is fitted at the solder side of the board. To make sure that they just about protrude from the front panel, the LED displays must be fitted at a height of 5.5 mm above the board surface (their own height is 6.5 mm). Stacked IC sockets, IC pin strips, or wire-wrap sockets are perfect for this.

Next, we start the construction of the insertion card. As already mentioned, this allows three different types of memory to be used. To begin with, fit surface-mounted (SMA) resistors R9 to R16. This is necessary because these tiny parts are very difficult to solder properly when the other parts are already on the board. Proceed by fitting all the smaller parts: resistors, capacitors, transistors and other low-profile parts. Note that the PCB allows for 100-nF IC decoupling capacitors with 5-mm or 7.5-mm pitch to be fitted. The quartz crystal is fitted horizontally on to the board surface. You may connect the quartz case to ground with the aid of a short length of stiff wire.

As shown in the introductory photograph, connectors K2, K4 and K5 are box headers. Either box header K4 or PCB connector K1 is fitted to leave you the choice between two types of printer output connector for PRINTER A: the PCB-mount 25-way sub-D connector gives a direct connection on the rear panel of the PC, while the 26-way box header allows a flatcable to be connected to another D-25 connector fitted somewhere else on the PC's rear panel. The choice is yours.



Think hard about your choice of the memory to be fitted in the printer buffer. At the time of writing this article, the most economical choice would appear to be a 1-MByte SIMM module. These modules may already be available as surplus items (or at least at reduced prices), since it is common practice these days to replace them with 4-Mbyte SIMMs. However, you may also have eight 1-Mbit ICs lying about (salvaged from an old XT?), which may be put to good use now. If discrete DRAM ICs are used, it is best to fit these in sockets. If a SIMM is used (1 or 4 MByte), you also require an appropriate socket. The pins of a SIPP module are best inserted into a row of pin sockets fitted at the track side of the board. This allows you to unplug the SIPP, and fit a SIMM later.

The front panel foil shown in Fig. 5 may be stuck on a 5¼-inch disk drive cover panel. Holes must be cut in this panel for the displays, the push-buttons and the LEDs. The front panel foil gives a professional touch to the printer buffer.

When the two printed circuit boards are completed, it is time to turn to the connecting cables. If you are satisfied with only one printer output, you only need to make the cable that links the insertion card to the display card. This 20-way cable has IDC sockets at both ends, and its length is geared to the distance between the two cards in the PC.

The two printers, PRINTER A and PRINTER B, occupy the same address (i.e., the address assigned to the printer buffer card), and they can be selected via the A/B push-button on the front panel. This allows, for instance, a label printer to be connected to the PRINTER A output, and a matrix or laser printer to the PRINTER B output.

Those of you who do not have enough space on the PC front panel may want to do without the push-button/display unit altogether. In that case, you will, however, forfeit the indications shown by the display, while it

Please note that this is the 1:1 component mounting plan of a double-sided through-plated printed circuit board. Owing to lack of space, the 1:1 track layouts of the component and solder side are not given as usual. Copies of this artwork (on paper only) are available free of charge from our editorial offices. Please send a stamped, addressed envelope, and quote 'PCB artwork 920009'. Overseas readers please send two International Reply Coupons (IRCs).

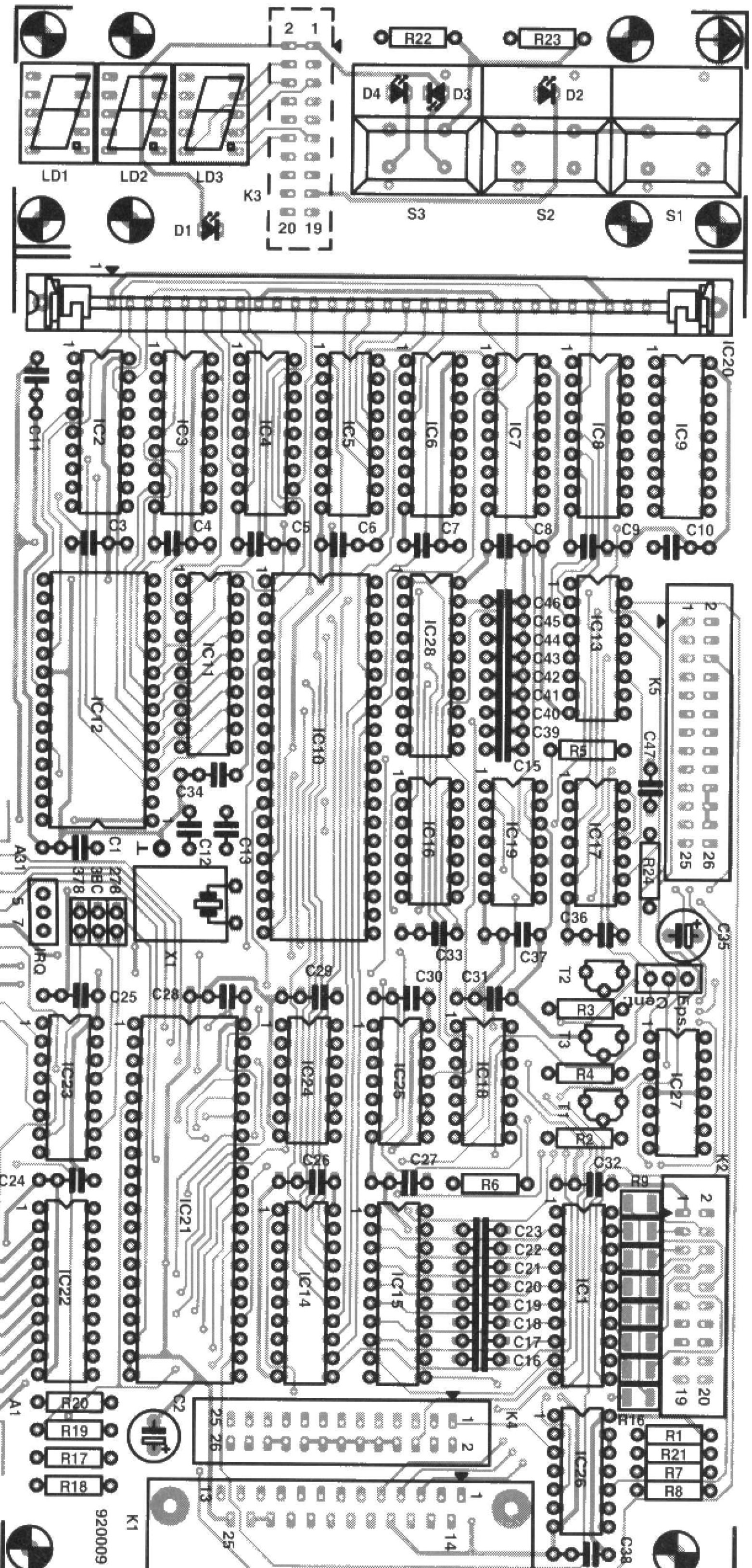


Fig. 4. Component mounting plan of the double-sided, through-plated printed circuit board for the printer buffer.

is still necessary to connect the control switches to K2. If you lack space, also consider building the display/controls unit into a small enclosure fitted with the front panel shown, and located near the printer.

Experienced constructors may want to replace two of the address selection jumpers, JP3, JP4 and JP5, by a toggle switch. Note, however, that this switch must not be operated when the PC is running.

Figure 6 shows the drilling details of the fixing bracket used to secure the insertion card to the rear panel of the PC. If you wish to create the second printer output, make a flatcable with a 26-way IDC socket at one end, and a 25-way sub-D connector (preferably IDC style also) at the other. From here on, it is all plain sailing. Simply untangle your ready-made IBM-to-Centronics printer cables, and connect the printers to the PC — one to the sub-D connector on the insertion card fixing bracket, the other to the 'external' sub-D connector. Label these outputs 'PRINTER A' and 'PRINTER B'.

Address selection

In case you have little experience in upgrading PCs, now is the time to unearth and consult the user manuals that came with your machine and its extension cards. Find out which address is occupied by the parallel printer port (LPT) already installed in the system. Alternatively, use a program like CheckIt to obtain this information. Next, choose a free LPT address, which will be $278, $378 or $3BC, and a suitable interrupt line, which will be IRQ5 or IRQ7. Do **not** use an address that is already occupied — this will cause awkward hardware conflicts. If necessary, use the DOS DEBUG utility to check the LPT address assignments. When using DEBUG to check, and if necessary, change, the LPT addresses, always keep in mind that you are working at the system software level, where mistakes can have serious, if not disastrous, consequences. Figure 10 lists a Pascal program that allows you to redirect the LPT address pointers much more comfortably than with DEBUG alone. Being able to 'move' the printer buffer card to LPT1 can be very useful in a number of cases, particularly with programs that insist on printing to a particular port (for instance, OrCad).

Communication type

The printer buffer card will behave either as a Centronics or an Epson interface. These standards differ marginally in respect of the handshake signal timing. The Epson mode is used by de-

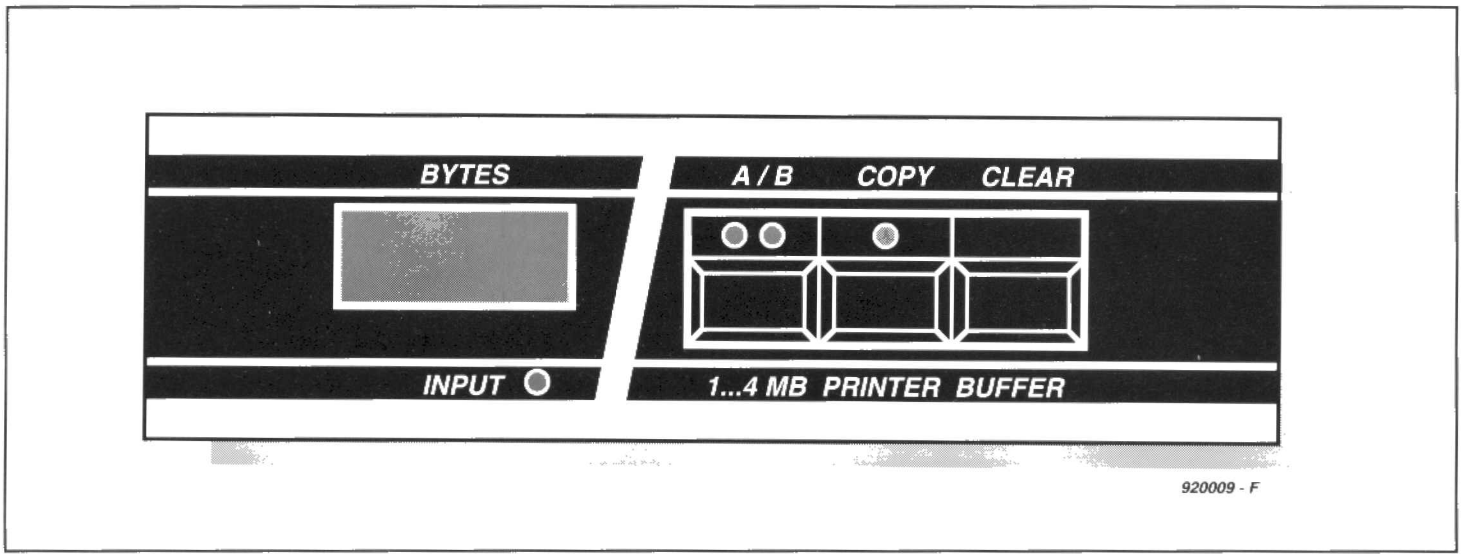


Fig. 5. Front panel layout.

fault. If necessary, set the 'Cent./Eps.' wire jumper on the printer buffer card as required.

Control software

The control software for the printer buffer insertion card resides in the system EPROM (order code ESS 6041), and is basically a modified version of the software developed for the '4-Mbyte printer buffer' described in the June 1992 issue of *Elektor Electronics*, with a few modifications to make it even more powerful. One of these modifications was necessary to enable the use of three-chip 1-MByte DRAM modules, which have a refresh organisation that differs from the 'older' eight or nine-chip modules. The flow chart of the control software, which has been written in MCS-51 assembler, is shown in Fig. 7.

At power-on, the printer buffer control software automatically 'measures' the size of the memory installed, and sends its findings to the displays for

COMPONENTS LIST

Resistors:

4	10kΩ	R1;R6;R18;R21
3	2kΩ2	R2;R3;R4
1	100Ω	R5
4	330Ω	R7;R8;R22;R23
8	330Ω SMA	R9-R16
2	1kΩ	R17;R20
1	33Ω	R19
1	1kΩ	R24

Capacitors:

24	100nF	C1;C3-C11; C24-C34;C36; C37;C38
2	100µF 16V radial	C2;C35
2	22pF	C12;C13
17	220pF	C15-C23; C39-C46
1	220pF	C47

Semiconductors:

1	LED, 3mm, red	D1
3	LED, 3mm (in keys S2 and S3; 2 × red, 1 × green)	D2;D3;D4
3	BC547B	T1;T2;T3
5	74HCT573	IC1;IC11;IC14; IC15;IC28
8	HYB511000 *	IC2-IC9
1	80C31BH-3 16P (16MHz)	IC10
1	EPROM 2764 (order code ESS6041; see page 70) +	IC12
2	74HCT138	IC13;IC23
2	74HCT00	IC16;IC18
1	74HCT04	IC17
2	74HCT74	IC19;IC27
1	SIMM or SIPP DRAM module; 1 MByte or 4 Mbyte (SIMM plus adaptor socket) *	IC20
1	UM82C11 (UMC) ++	IC21
1	74HCT245	IC22
1	74HCT10	IC24
1	74HCT27	IC25
1	74HCT4066	IC26
3	HD1105G (Siemens) +++	LD1;LD2;LD3

* See text for memory configuration options. SIMM/SIPP and discrete ICs can NOT be combined.
+ Supplied with printed circuit board 920009.
++ UMC (United Microcomponents Corp.) product. Suggested suppliers:
(1). Manhattan Skyline Ltd., Manhattan House, Unit 1, The Switchback, Gardner Road, Maidenhead, Berks SL6 7RJ. Tel. (0628) 778686. Fax: (0628) 782812.
(2). C-I Electronics, P.O Box 22089, 6360 AB Nuth, Holland. Fax: +31 45 241788.
+++ UK distributor: ElectroValue, 28 St. Jude's Road, Englefield Green, Egham, Surrey TW20 0HB. Tel. (0784) 433603. Fax: (0784) 435216.

Miscellaneous:

1	PCB-mount 25-way sub-D socket	K1
2	20-way box header	K2;K3
2	26-way box header	K4;K5
1	Digitast * push-button with 17-mm cap	S1
1	Digitast * push-button with 17-mm cap and built-in LED (=D2)	S2
1	Digitast * push-button with 17-mm cap; locking action and two built-in LEDs (red and green; D3 and D4)	S3
1	16MHz quartz crystal	X1
1	20-way IDC socket	
1	25-way IDC sub-D socket	
1	26-way IDC socket	
1	Printed circuit board 920009 (see page 70)	
1	Front panel foil 920009-F (see page 70)	
1	Approx. 50cm long piece of 36-way flatcable	
1	PC insertion card support bracket, e.g., Fischer** Type XTN25SL-U	

* ITT Cannon UK, Jays Close, Viables Estate, Basingstoke, Hants RG22 4BW. Tel. (0256) 473171. Fax: (0256) 23356.
** Dau Components Ltd., 70-74 Barnham Road, Barnham, W. Sussex PO22 0ES. Tel. (02435) 53031. Fax (02435) 553860.

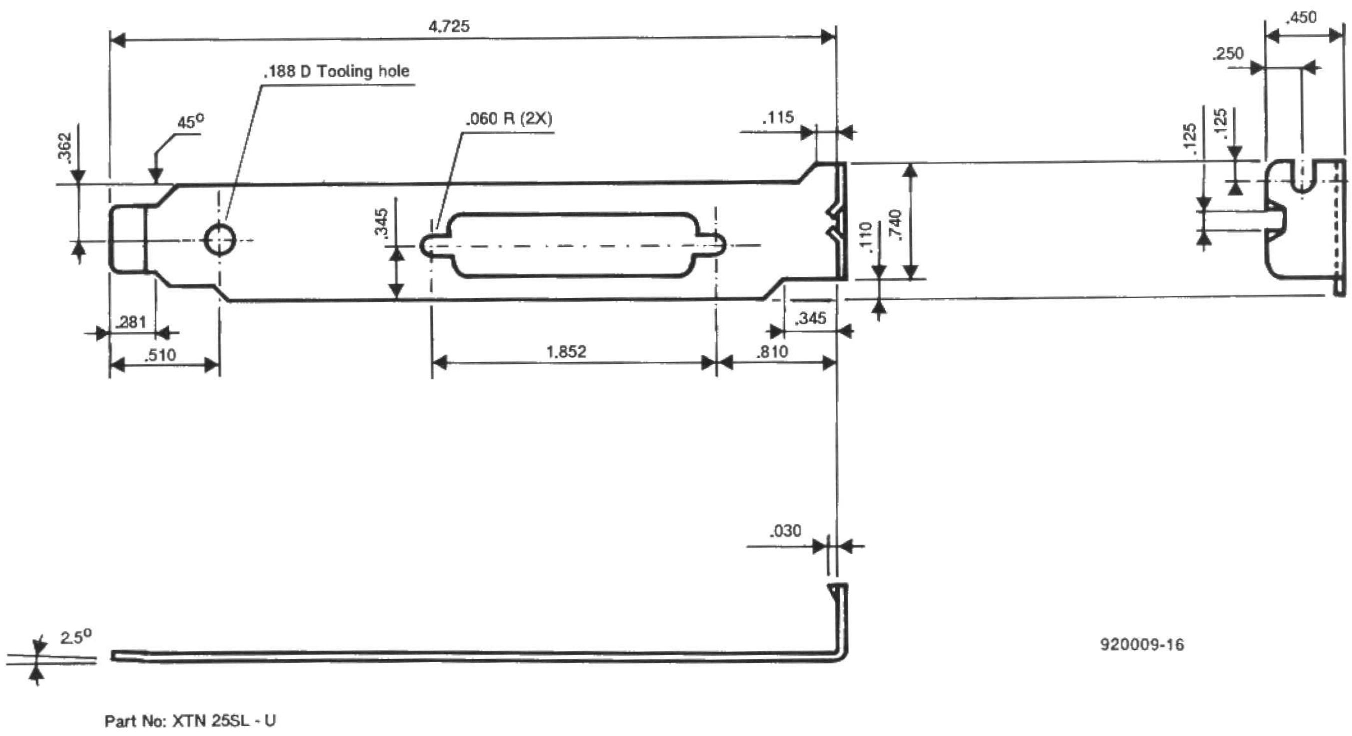


Fig. 6. Insertion card fixing plate dimensions and drilling details.

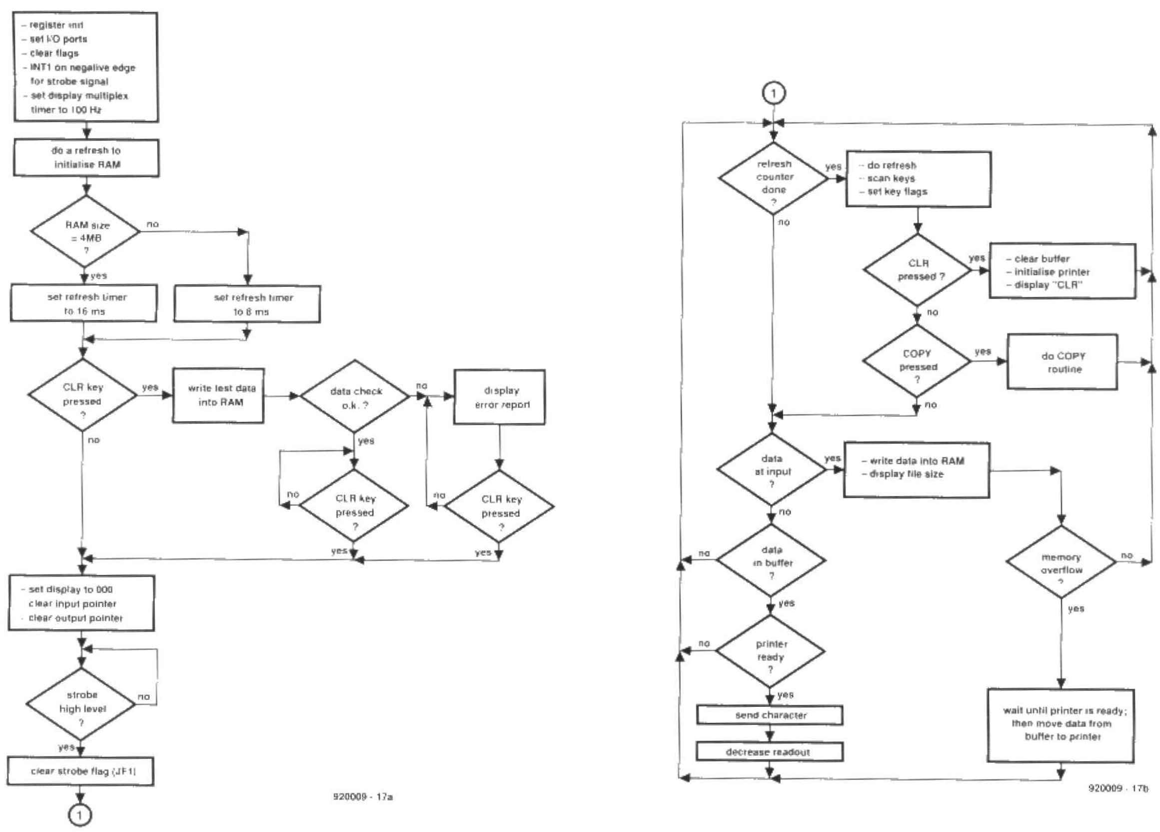


Fig. 7. Control program flowchart.

```
C:\>debug          COM1    COM2    COM3    COM4     LPT1    LPT2    LPT3
-d 0:400
0000:0400  F8 03 F8 02 00 00 00 00-78 02 BC 03 00 00 F1 4B   .........x......K
0000:0410  63 44 BF 80 02 00 00 00-00 00 30 00 30 00 0D 1C   cD........0.0...
0000:0420  64 20 20 39 30 0B 3A 27-34 05 30 0B 30 0B 0D 1C   d  90.:'4.0.0...
0000:0430  73 1F 0D 1C 64 20 65 12-62 30 75 16 67 22 00 00   s...d e.b0u.g"..
0000:0440  6F 00 C0 00 00 00 00 00-00 03 50 00 00 10 00 00   o.........P.....
0000:0450  00 08 00 00 00 00 00 00-00 00 00 00 00 00 00 00   ................
0000:0460  07 06 00 D4 03 29 20 88-05 87 90 00 8E EC 0D 00   .....) .........
0000:0470  00 00 00 00 00 01 00 00-14 14 14 14 01 01 01 01   ................
-q

C:\>setlpt 378 278 3bc

C:\>debug                                        LPT1    LPT2    LPT3
-d 0:400
0000:0400  F8 03 F8 02 00 00 00 00-78 03 78 02 BC 03 F1 4B   .........x.x....K
0000:0410  63 44 BF 80 02 00 00 00-00 00 36 00 36 00 62 30   cD.........6.6.b0
0000:0420  75 16 67 22 0D 1C 64 20-20 39 30 0B 3A 27 34 05   u.g"..d  90.:'4.
0000:0430  30 0B 30 0B 0D 1C 63 2E-0D 1C 64 20 65 12 00 00   0.0...c...d e...
0000:0440  29 00 C0 00 00 00 00 00-00 03 50 00 00 10 00 00   ).........P.....
0000:0450  00 16 00 00 00 00 00 00-00 00 00 00 00 00 00 00   ................
0000:0460  07 06 00 D4 03 29 20 88-05 87 90 00 D4 EE 0D 00   .....) .........
0000:0470  00 00 00 00 00 01 00 00-14 14 14 14 01 01 01 01   ................
-
```

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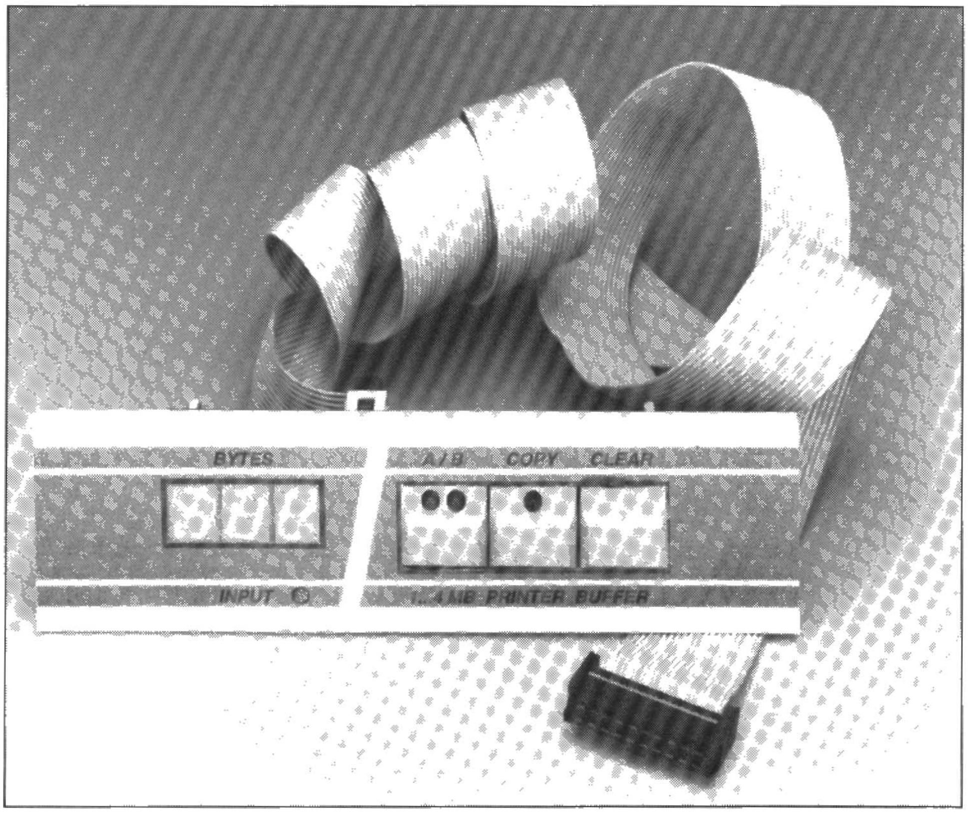
Fig. 8. Screendump illustrating how the DOS DEBUG utility can be used to check, and SETLPT (below) to change, the three LPT I/O addresses in the PC (LPT = line printer). Note that the parameters for SETLPT are [LPT1 address] [LPT2 address] [LPT3 address] in that order. If necessary, SETLPT (with parameters) may be put into the AUTOEXEC.BAT file.

you to read. The indications '1.00' and '4.00' mean that 1 MByte or 4 MByte of RAM have been found, respectively.

On completion of the RAM test, the display is cleared, and indicates '000'. Next, the printer buffer actuates the INIT line of the printer to make sure that this is properly initialized. Also, the state of the PC's STROBE line is checked. When the line goes high, the buffer switches the printer to receive mode.

Apart from the memory size check at power-on, the control software also provides an extensive RAM test, which is particularly useful when a fault is suspected in the memory. The extensive RAM test is started by keeping the CLEAR push-button pressed for some time after the PC's reset signal has disappeared. The RAM test writes bytes to the memory, and reads them back to check the integrity. Provided there are no faulty RAM locations, the display can be seen to count up to '1.02' (1,024 Kbytes) or '4.09' (4,096 Kbytes) for the 1 MByte or 4-MByte version respectively. The RAM size check speed is about 1 MByte per minute, so you will need a bit of patience for the larger version. The indication 'Err' appears when an error is found. If this happens, there is probably no other option than to replace the entire memory, be this a module or eight ICs. If an error is found (i.e., the byte read back is not equal to the byte written), the faulty address block (in 'Kbytes') flashes on the display. You can also see which bit(s) is (are) at fault, because the bits written are indicated on the left-hand display, and the bits read back, on the right-hand display (see Fig. 9). The display segments



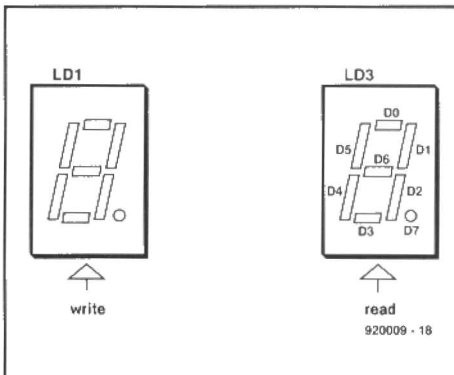
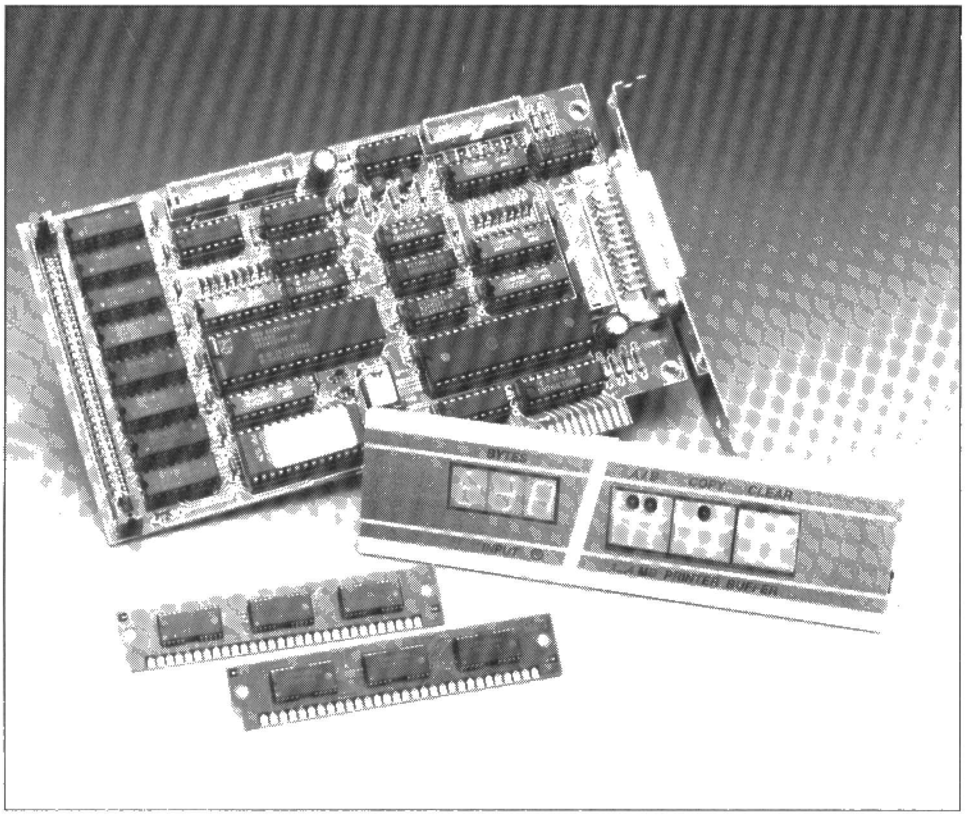


Fig. 9. Correspondence between the LED display segments and bits in the datawords contained in the buffer memory.



and the decimal points correspond to the databits D0-D7 as indicated in the drawing. The test will remain 'stuck' at the faulty RAM location. However, pressing CLEAR again allows you end the memory test at any time.

As already mentioned, the DRAM refresh operation is implemented in software rather than hardware (which is more usual). Here, the microcontroller's PSEN\ line is used for the purpose. PSEN\ goes low two times every cycle, and serves to generate the RAS\ (row address select) signal for the DRAMs. This is achieved by having the microcontroller jump to a routine that contains 512 2-cycle 'dummy' instructions that do nothing except increment address lines A0 to A9. This type of DRAM refresh is generally referred to as 'RAS-only refresh'. The read/write operations to and from the RAM are under the control of the WR\ (write) line of the processor. To address a RAM location, the processor first has to set up the column address, and then the row address. To do this, we start by making the RAS\ line high via the microcontroller's T1 output (RAS enable, pin 15). Next, the row address is placed on the data bus with the aid of a MOVX instruction. Simultaneously, the WR\ line is actuated, which causes the output of bistable IC19a (pin 6) to go low. Via gates IC166 and IC16c, this signal arrives at the RAS\ inputs of the DRAM ICs.

At this stage, we require only one more thing: the column address. This time, we start by actuating the CAS-enable signal. This is achieved by actuating the T0 output of the microcontroller (pin 14). Next, the column address is placed on the data bus with the aid of a MOVX instruction. As with the row addressing, WR\ is made low, and arrives at the CAS\ inputs of the DRAM ICs via gates IC16a and IC16d.

Since the output of bistable IC19b has been set by the rising edge of the WR\ signal, the other bistable, IC19a, is reset, and, consequently, the RAS\ line also. An *R-C* network, R5-C15, causes IC19b to be reset a little later. At this stage, the microcontroller can have access to the memory, i.e., bytes can be read or written.

Practical use

The practical operation of the printer buffer is pretty simple, hence you will be relieved to hear that there is no hefty manual for a change. Simply insert the card is a free slot, and connect your printer(s) to the buffer output(s). After switching on, you will see the display 'counting' RAM locations.

Once the display indicates '000', the buffer is ready for use. Before directing a print file to the buffer, make sure that the printer is switched on, loaded with paper, and 'on line'. Send the file to the printer buffer (i.e., to the selected LPT), and you will see the INPUT LED light, and the display reading increasing up to a value roughly equal to the size of the file. Next, the display reading decreases again, which indicates that the buffer is feeding its data to the printer.

Pressing the **CLEAR** key allows you to clear the buffer memory. The CLEAR key has a second function, which is described below.

```
program setlpt;

uses dos;

var

printer1, printer2, printer3: word;

procedure parameters;

var

data1, data2, data3: string;
i,parcount: integer;

begin
      data1:='0';
      data2:='0';
      data3:='0';
      parcount:= paramcount;
      if parcount >=3 then data3:=paramstr(3);
      val ('$'+data3,printer3,i);
      if parcount >=2 then data2:=paramstr(2);
      val ('$'+data2,printer2,i);
      if parcount >=1 then data1:=paramstr(1);
      val ('$'+data1,printer1,i);
end;

begin
  parameters;
  if printer1<>0 then
    begin
      mem[0:$408]:=lo (printer1);
      mem[0:$409]:=hi (printer1);
    end;
  if printer2<>0 then
    begin
      mem[0:$40A]:=lo (printer2);
      mem[0:$40B]:=hi (printer2);
    end;
  if printer3<>0 then
    begin
      mem[0:$40C]:=lo (printer3);
      mem[0:$40D]:=hi (printer3);
    end;
end.
```

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Fig. 10. A small Pascal program to manipulate the parallel printer port address assignments.

The **COPY** key is used to send another copy of the file to the printer. Only in COPY mode, pressing **CLEAR** briefly causes the printing process to stop. Pressing COPY again continues the printing from where it was interrupted, but only for the last file sent to the buffer.

Pressing **CLEAR** two times clears the memory, and initializes the printer. ■

Correction

Please note that the fax number of C-I Electronics given in the Components List on page 56 of this article is incorrect. The correct number is **(+31) 45 241877**.

DX TELEVISION

A two-monthly column by Keith Hamer and Garry Smith

By the middle of September 1992, the number of Sporadic-E openings dwindled dramatically, indicating that the main season was almost over. Mid-October saw the reappearance of activity with several days of sustained openings from many European countries.

November and December were quiet months for Sporadic-E signals but tropospheric DX-ing was possible in Band III and at UHF on several dates. Signals from Belgium, the Netherlands, Luxembourg, Germany, Austria and Switzerland were monitored on November 8th. The long-awaited F2-layer activity was minimal with only a few pictures noted on the lower Band I channels, probably from Thailand or the Middle East. Meteor-scatter DX produced some prolonged 'pings' of video, especially at the beginning of January 1993.

Tropospheric reception

Perhaps at this stage a few notes on tropospheric reception would not go amiss, especially for newcomers to long-distance television (DX-TV) who may be reading this new column in *Elektor Electronics* for the first time.

Tropospheric reception is associated with anticyclonic weather conditions (high pressure systems). These can occur at any time of the year and may exist for several days. Unlike Sporadic-E propagation, reception is relatively stable, even over several hours, although there is a noticeable drop-off in activity between midmorning and early evening. A skip-distance is not involved with this type of propagation, so reception distances can extend from semi-local ITV transmitters to European ones up to 800 km away.

If, while tuning through the UHF band, you notice other UK stations which are not normally present, then this is the time to be on your guard. A check should also be made for signals in Band III. Band I is not as readily affected by tropospheric propagation. In central parts of the United Kingdom, signals from Dutch, Belgian, German and French stations may also be noticed in addition to the extra BBC and ITV channels. During really intense openings, Czechoslovakian and Polish transmissions may emerge. DX-ers living in the north-east of the UK may first experience Danish and north German signals; indeed these seem to show on a regular basis. In East Anglia and Kent, Benelux transmissions are present to a degree all the while, ranging from poor to excellent quality in terms of signal strength. These stations can be used to supplement UK viewing if a converter is used, since many European stations use VHF channels as well as UHF. The converter will also take care of the sound carrier differences because the sound carriers of foreign transmissions are not compatible with those of the UK.

Reception reports

The prolonged foggy conditions during December produced some ideal tropospheric reception conditions. Several announcements made over the UK networks warned that interference to television pictures would occur. In some locations, foreign pictures actually swamped the UK transmissions.

Simon Hamer (Powys, Wales) went mobile DX-ing into the Welsh mountains with temperatures as low as −10 degrees Centigrade. Despite the icy conditions, he was rewarded with some startling catches including the Faröes, Finland and the new Norwegian TV-2 network, all in Band III. Polish, Czechoslovakian and Austrian UHF reception was also noted by Simon.

Many other DX-ers noted transmissions from the Benelux countries, Germany and Denmark. The latter country, with its network of high-power UHF transmitters, virtually guarantees reception in the UK during enhanced tropospheric conditions!

VHF/UHF DX-TV converter

During the period under review, the authors were able to evaluate the performance of the Amstrad MP3 VHF/UHF receiver which was originally designed for converting a computer monitor into a TV set. These units are manually tuned and have an in-built loudspeaker. For DX-TV use it is necessary to fit audio and video take-off sockets for hooking up to the TV via a video recorder, unless you use a modulator. The unit handles teletext signals quite well — it was quite a novelty being able to access the Dutch and Belgian teletext pages in Band III. One of these receivers has now been modified to incorporate variable vision IF bandwidth to boost the receiving threshold when monitoring extremely weak signals. Video polarity selection has also been developed by one of the authors so that standard, or French, video can be selected.

If you see these units advertised on the surplus market, make sure that a slide switch is present at the front of the unit for band selection. There are some UHF-only units around, so watch out! Some of the VHF/UHF units already have the 5.5 MHz sound crystals fitted for Continental sound. *HS Publications* can supply ready modified units for DX-ing, some with variable IF bandwidth and dual UK/Continental sound. Availability and prices may be obtained upon request by sending an SAE (or 2 IRCs) to: HS Publications, 7 Epping Close, Derby DE3 4NP, England.

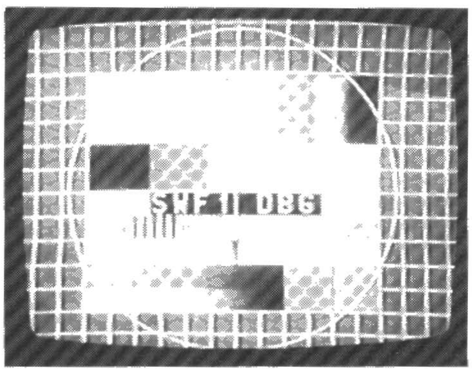

Fig. 1. A German test pattern from the Dillberg transmitter in the Südwestfunk broadcasting region.

Reception Logs

Now to the reception log for December and early January. Many thanks to the following enthusiasts who have contributed reports: Bob Brooks, Andrew Jackson, Stephen Michie, Peter Chalkley, Simon Hamer, Ian Johnson.

December 1992 Log

06.12.92: CIS channel R2 via MS (Meteor-Scatter); Denmark channel E4 via MS.

10.12.92: Norway, Denmark and Sweden (all Band I signals) via MS. Spain (Band III/UHF) via tropospheric propagation.

14.12.92: Belgium (UHF); Germany (Band III/UHF).

21.12.92: Denmark (Band III/UHF); Sweden Band III/UHF); the Faröes (Band III); Norway (Band III); Finland (Band III); Poland (Band III).

24.12.92: CIS (Band I) via MS.

28.12.92: Austria, Poland, Czechoslovakia and Switzerland (all in Band III and at UHF) via tropospheric propagation.

January 1993 Log

02.01.93: Spain on channel E3 via MS.

04.01.93: Spain E2; Denmark E3; Sweden E3 (all via MS).

06.01.93: Denmark F4 via MS.

Peter Chalkley tells us that during last summer he has been using the RSGB Sporadic-E Hotline run by Jim Bacon, which provides an accurate forecast. The number is (0426) 952211.

Service information

Germany: new transmitters in operation:
Wiederau to relay ARD E9, ZDF E42 and MDR E22; MDR E52 with regional programmes for Sachsen-Anhalt; ARD E35 with regional programmes for Sachsen-Anhalt
Halle: SAT-1 E57
Magdeburg: SAT-1 E47
Düsseldorf: VOX E39 (formerly used by WDR-3)
Wesel: VOX E59 (formerly used by WDR-3)

RTL Plus plans to launch a second channel in the not too distant future.

Czechoslovakia: following the formation of two independent states on January 1st 1993, each of the republics will have its own radio and TV networks. These are:
- Czech Republic: First Programme (CTV 1) via the former CTV channels;
- Second Programme (CTV 2) via the former F1 channels;
- Third Programme (TA3 — formerly OK-3) via existing channels.

A private TV company called Premiera has obtained a licence for regional broadcasts in Prague and central Bohemia. The new service will use channel R24. A regional TV studio has been set up in Decín and broadcasts twice a week via certain transmitters.

Slovak Republic: First Programme (STV 1) via the former F1 channels, and second Programme (STV 2) via the former STV channels.

Videotelevizía, a regional TV station in Prievidza, is operating on channel R46 with 50 kW ERP.

The PAL colour system is now used throughout the Czech and Slovak Republics, except networks using the old F1 transmitters. Most of these transmitters are technically outdated and cannot meet the more stringent performance requirements of the PAL system. An immediate switch would result in a picture quality inferior to that of the current SECAM transmission.

In the Czech and Slovak republics the changeover to PAL will eliminate the PAL/SECAM output transcoding which is currently employed — this being a significant source of picture degradation.

A change from PAL D/K (6.5 MHz sound) to PAL B/G (5.5 MHz sound) is also under investigation. One of the technical problems that will be encountered is the reduction in bandwidth of the video channel. The first transmitter to use the B/G standard will be located near Bratislava.

Russia: an experimental teletext service called 'Teleinf' is being broadcast via the OK-1 network. Initial experiments used Italian equipment with the page header in Italian.

Poland: TVP is expected to change from SECAM to PAL in the near future. Already certain programmes are aired in PAL colour.

Rumania: there are plans to privatize the TVR-2 network. An independent and commercial TV service called 'Canal 2' is expected to start broadcasting soon. In addition, at least 13 local TV services are operational throughout Rumania.

TVR-2 is relaying 'Mesager', the Moldavian TV news, from 1845-1930 UTC daily, followed by a selection of satellite material including TVE, TV5 and BBC World Service Television until closedown.

Moldava: the Rumanian TVR-l programmes are now relayed over the ex-Moscow channels. The Band I transmitters are Cahul R1, Bender R1 and Cimislia R2.

The interchange of programmes between the two countries may create identification problems for the DX-er unless the colour signal is received. Moldavia uses SECAM, Rumania uses PAL.

Norway: a second network commenced on September 5th 1992, based at Bergen. The network is financed by advertising.

Netherlands: AFN-TV at Soesterberg has moved from channel E72 to E24 with vertical polarisation. There are plans to install a 50-MHz beacon in the centre of Rotterdam.

Switzerland: the Swiss Pay-TV film channel (on channel E69) Télécíné Romandie (TCR) is to close due to financial difficulties, even after so many years of operation. The owners have another Pay-TV project on the drawing-board with French-speaking movies via satellite at a reasonable price.

The Swiss PTT have studied the possibility of a fourth terrestrial TV channel. Unfortunately the lack of spare frequencies close to the borders of Italy, Germany and France means that not everyone would receive it.

France: 'France 3' may soon be broadcast by satellite via Telecom B. There are some viewers in remote areas who cannot receive terrestrial transmissions and rely on satellite broadcasts. Some of the private UHF transmitters close to the Swiss border (which relayed 'La Cinq') have been relaying 'Canal J' and 'Canal Jimmy'. In the future 'Canal Plus' may be relayed because several Swiss cable networks are distributing the former services.

Spain: huge job losses are on the cards at RTVE. Advertising revenue has plummeted and many of the advertising slots simply show trailers of forthcoming programmes.

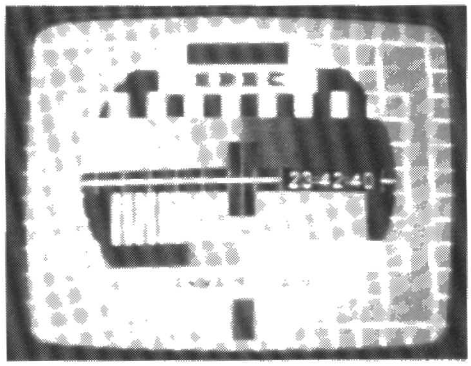

Fig. 2. Band III troposheric reception from Denmark.

Problems began when competition from other services began to take hold. In order to compete, RTVE virtually doubled its broadcasting hours. Currently 'Tele 5' and 'Antena 3' have taken the lead in primetime ratings.

United Kingdom: The BBC's Ceefax teletext service has been revamped offering a 'faster' service. Viewers have a choice of 'SPEED' on BBC-1 with summaries of news and other items, and 'DEPTH' on BBC-2 where topics are covered in greater detail.

Since January 1st 1993, the Oracle teletext service on ITV and Channel 4 has been replaced by 'Teletext on 3' and 'Teletext on 4' respectively. The new company provides detailed regional news coverage via Channel 4.

The name of the region (for example, Central TV or Anglia TV) is now displayed at the foot of the picture for a few seconds after selecting a different channel. This will greatly assist European DX-ers in identifying the actual UK transmitter.

Belgium: due to financial problems, the TV transmitter at Oostvleteren has been taken out of service. Viewers must now retune to the main 1000-kW transmitter at Egem (TV-1 E43, TV-2 E46).

Denmark: The DR transmitter at Fyn (channel E3) has increased its power to 25 kW. The other Band I outlet at København (channel E4) remains at 50 kW.

This month's Service Information was kindly supplied by Gösta van der Linden and the Benelux DX Club, Netherlands; Bernd Trutenau, Lithuania; Pertti Salonen, Finland; Reflexion Club, Germany; Roger Bunney, UK; Jurgen Klassen, Germany; Munteanu Corneliu, Rumania; Simon Hamer, UK; André Gille, France. ❑

Please send any news about DX-TV in your part of the world to: Keith Hamer, 7 Epping Close, Derby DE3 4HR, England.

SIMPLE, LOW-COST ANTENNA TEST INSTRUMENTS - 2

By Joseph J. Carr

Impedance Bridges

We can make antenna impedance measurements using a variant of the old-fashioned Wheatstone bridge. Figure 8a shows the basic form of the bridge in its most generalized form. The current flowing in the meter will be zero when $(Z_1/Z_2) = (Z_3/Z_4)$. If one arm of the bridge is the antenna impedance, and two other arms are fixed, then we can adjust the remaining arm to null the bridge and thereby make the measurement. For example, a common tactic is to make Z_1 and Z_3 fixed resistors with a value close to the middle of the range of impedances being measured. If Z_2 is a variable resistance, the resistance that balances the bridge, as indicated by the null condition on the detector, is equal to the resistive component of the unknown impedance.

A typical example is shown in Fig. 8b. The antenna, or other unknown impedance, is connected to J2 and forms one arm of the bridge. The second arm of the bridge is a fixed resistance (R2). The value of R2 should be 50 Ω or 75 Ω, depending upon the value of the expected antenna impedance. The choice of 68 Ω is a good compromise for meters to operate on both types of load. Resistor R2 must be a carbon composition, metal film or other non-inductive type. Wirewound resistors are inductive and at RF frequencies they act like tuning coils rather than resistors.

The other two arms of the bridge are the reactances of C1A and C1B, which is a single differential capacitor. A differential capacitor (Fig. 8c) is one that has two sections that are mechanically displaced relative to each other by 180 degrees, and sharing a common rotor shaft. Both sections have the same capacitance, but the design makes one section at maximum capacitance while the other is at minimum capacitance. Thus, as the shaft is turned, the capacitance of one section decreases the same amount as the capacitance of the other section increases.

It is possible to make an impromptu differential capacitor by mechanically ganging two identical variable capacitors. The kind to use are straight line capacitance air variables that have the rotor shaft protruding out the back as well as the front. Use a shaft coupler to connect the rear shaft of one capacitor to the front shaft of the other. Set one of the capacitors to maximum capacitance (plates fully meshed) and the other to minimum capacitance (plates fully unmeshed). It is not as good as a real differential capacitor, but works well enough if done properly.

The detector in this circuit is a germanium signal diode (1N34 or 1N60 are suitable). Resistor R3 forms a load for the diode, and a microammeter (M1) serves as a detector display. Typically, 0 to 100 μA meters are used. An RF choke (RFC1) is used to prevent RF from interfering with the meter movement.

Connect an unknown impedance to J2, and a signal source at the desired fre-

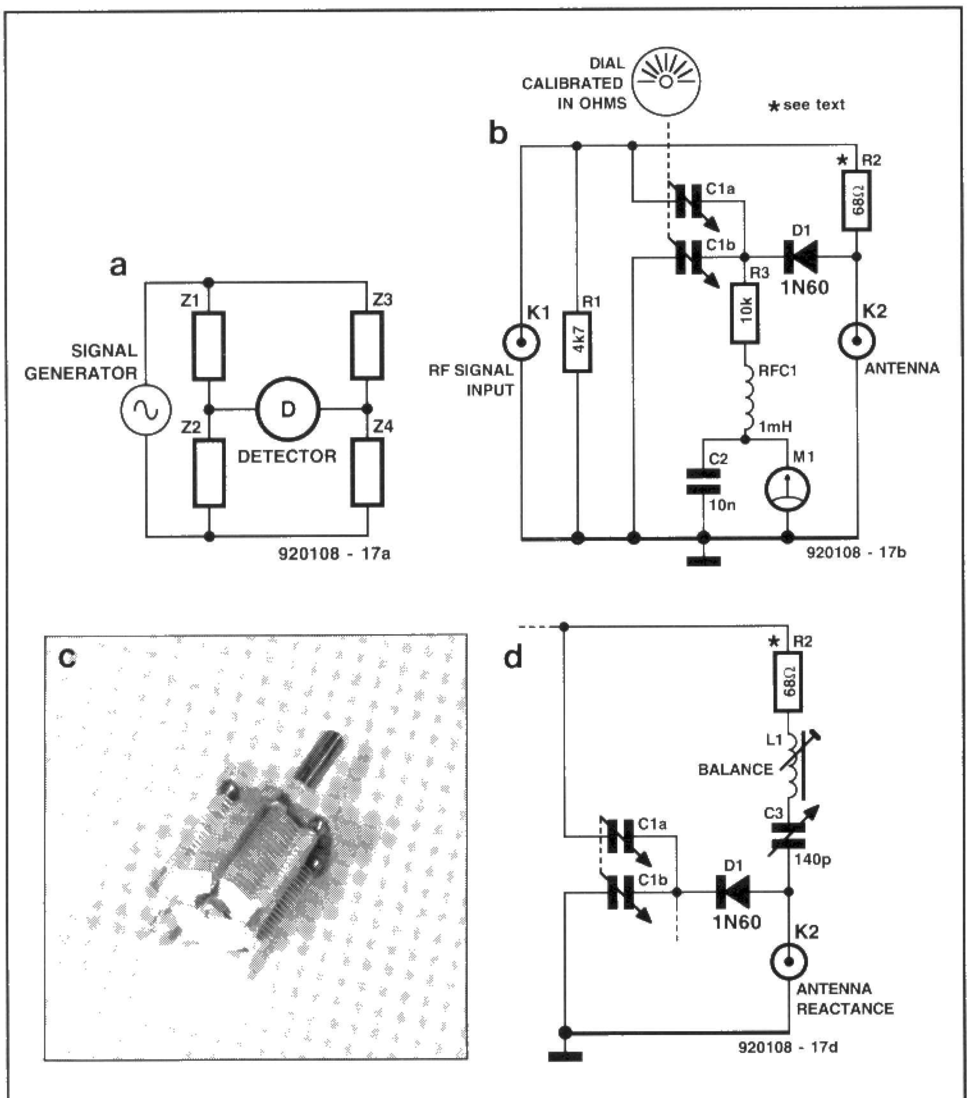

Fig. 8. a) Generic bridge circuit; b) RF bridge based on a differential capacitor; c) a 2 × 150 pF differential variable capacitor; d) Modification of circuit to provide BALANCE and REACTANCE controls.

Fig. 9. Commercial impedance bridge for amateur and hobbyist use.

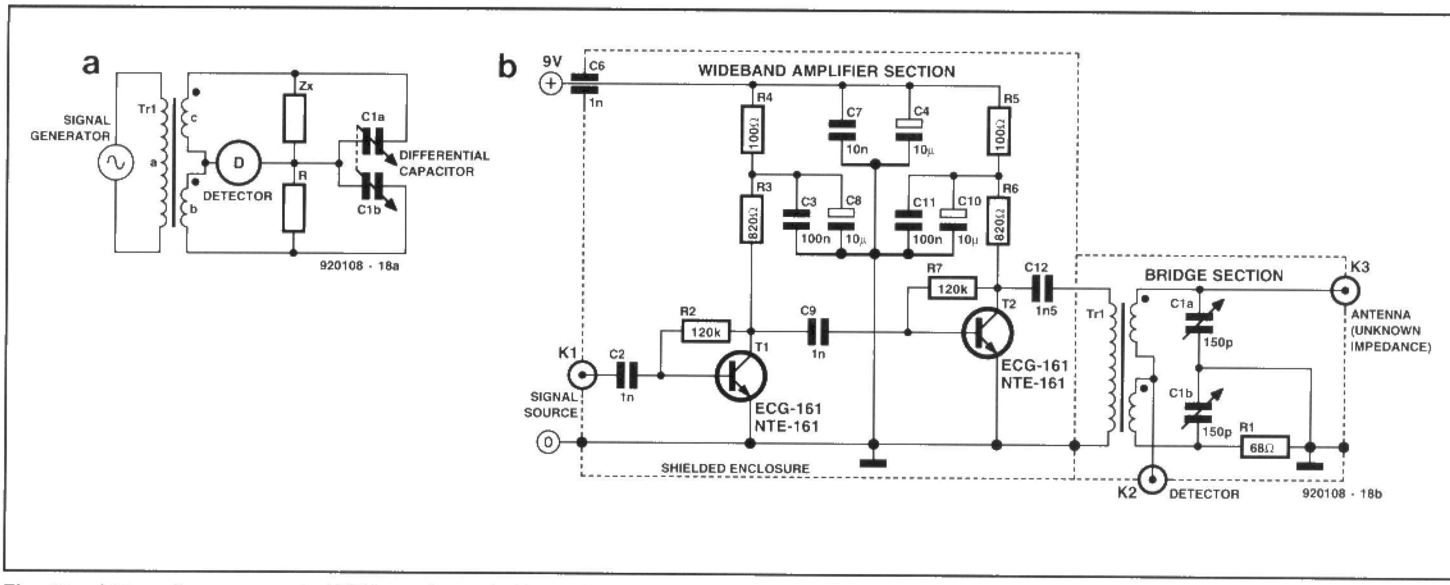

Fig. 10. a) Transformer coupled RF impedance bridge; b) practical circuit using RF amplifier ahead of bridge.

quency to J_1, and then tune differential capacitor C_1 until the meter indicates a null; read the antenna impedance from the dial ganged to the shaft of C_1. At least one instrument allowed the technician to plug in a resistor element equal to system impedance.

Calibration of the instrument is simple. Various noninductive, carbon composition resistors, having standard values from 10 Ω to 1000 Ω, are connected one at a time across antenna jack J_2 (use very short leads). The bridge is then nulled, and the value of the load resistor is inscribed on the dial at the point. Keep in mind that there is variation in the value of resistance (which is why tolerance colour bands are used), and there is some random variation in the actual measurement process. If you want a more accurate dial, either use 1-percent tolerance resistors, or hand select resistors for closeness to the design value by using an ohmmeter. Make the measurement at each resistance value five to ten times, and take the central point in their spread as the actual point where that resistance is to be marked.

The basic circuit of Fig. 8b is useful only to measure the resistive component of impedance. We can modify the circuit as shown in Fig. 8d to account for the reactive component. In this particular case, an inductive reactance, L_1, is placed in series with a variable capacitor, C_3, at the antenna jack (J_2). The capacitor is used to measure the reactive component of the impedance. The inductor (L_1) serves as a balance control. To adjust L_1, first null the bridge using C_1 (it will be a shallow null until L_1 is adjusted correctly). Set C_3 to midscale (70 pF) capacitance, and then adjust L_1 to produce a null. The value of L_1 depends on the operating frequency, and is designed to series resonate with C_3 at that frequency. Some impedance bridge projects based on this principle have used plug-in coils for L_1, allowing a different inductor to be used for each band. The explanation for not using a switch selectable inductor is that the stray capacitances and inductances inherent in switching make the bridge difficult to null. I suspect that modern switch methods, especially those methods based on PIN diodes, would render that argument less viable. Also, use of small DIP relays could solve the 'strays' problem.

A problem with the circuit of Fig. 8 is that it requires a fairly high signal level to get a decent deflection of the meter movement. This means that ham operators can use it with ease, but SWLs and others who are not licensed to use transmitters would find it a bit difficult to use. The solution is to use either an RF amplifier ahead of the bridge, to amplify the low-level output of the typical RF signal generator, or a DC amplifier ahead of the meter movement. I have seen projects where both forms of amplification were used.

An example of a commercial handheld radio antenna impedance bridge is shown in Fig. 9. This instrument uses the capacitor-resistor bridge circuit, and has an internal wideband RF amplifier to allow the use of low-level signal sources such as a workbench signal generator. A control switch cuts the amplifier in or out of the circuit, depending on whether or not it is needed.

A variant of the impedance bridge circuit is shown in Fig. 10a. In this circuit, the signal source is connected to the bridge via a transformer, Tr_1. These transformers are typically wound on a toroidal form, using the trifilar winding technique to ensure close coupling of all three windings. The dots at the ends of the two secondary windings (Tr_{1B} and Tr_{1C}) are used to indicate the same end of the two windings.

Figure 10b shows a version of this circuit that was built and tested by this author for this article (actual prototype in Fig. 10c); it is designed to operate in the high frequency (HF) shortwave portion of the radio spectrum, and produced good nulls into a 50-Ω dummy load at frequencies as high as 32 MHz. There is a wideband RF amplifier provided in the circuit to permit the use of a relatively low-level RF signal generator.

The bridge circuit shown is adapted from Fig. 10a. This version is designed to operate with coaxial cable transmission lines, which is commonly used by SWLs and ham operators alike. The bridge elements include the sections of the differential capacitor (C_{1A} and C_{1B}), the unknown antenna impedance and a fixed resistor (68 Ω). As before, the resistor must be a carbon composition or metal film resistor.

A three-winding transformer (Tr_1) drives the bridge circuit. This transformer is wound in the trifilar manner, using 10 turns of 26 AWG (approx. 0.5 mm dia.) enamelled wire, on a T37-RED toroidal core. A little trick can be used to make trifilar winding easy, if you are careful. Cut three strands of 26 AWG enamelled wire, each about 75 cm long, and lay them parallel to one another. Tie one end of the wires together and mount it in the chuck of a hand drill (a variable speed electric drill also works, but it must be run very slowly). Clamp the other end in a bench vise. Operate the drill to slowly twist the wires together until there are about six twists per centimetre (not terribly critical). Wear safety goggles or glasses when doing this job.

The detector can be any device or instrument capable of detecting the HF signal. I used both an oscilloscope (with 50 MHz –3 dB bandwidth) and a shortwave radio receiver (equipped with a signal strength or 'S-meter') as detectors. One could also use the passive analogue meter detector shown earlier in Fig. 10.

The bridge circuit must be built inside of a shielded enclosure to prevent interaction between the external environment and the bridge components. The bridge

compartment was built inside a larger shielded box that contained the bridge and a wideband RF amplifier.

The wideband amplifier consists of a pair of VHF/UHF npn transistor stages in cascade. The transistors used were easily available television service 'universal' replacements. I used the NTE-161 type (because of easy local availability), but it is the same as the ECG-161, GE-39 and SK-3716. For those people who do not have these replacement lines available, try any general-purpose n-p-n VHF/UHF RF amplifier transistor in the same class ($F_t \approx 800$ MHz, $H_{fe} \approx 60$, P_d= 200 mW, I_c=50 mA).

All of the capacitors in the circuit, except for those marked with polarity symbols and C6, are disk ceramics. Because this amplifier is wideband, use a good grade of capacitor, and trim off the bit of ceramic package material that usually flows down the leads so that there is only a minimum of excess lead left. A small needle nose pliers will usually crush the offending ceramic flow to allow its removal.

The polarized capacitors in the circuit can be either tantalum or aluminium electrolytics. Although most experts seem to prefer the tantalum types in applications such as Fig. 10b, I experienced no difficulty at all with even the cheaper grades of radial lead aluminum electrolytics actually used in the construction of the circuit.

Capacitor C6 is a 1000-pF ceramic feedthrough capacitor. It mounts on the rear panel, and allows DC current to be passed through the panel while bypassing RF signals to ground. These capacitors are not easy to locate, and may not be strictly necessary if the lead length between C7 and the point where the power lead passes through the panel is kept very short. Good design practice, however, suggests the use of the feedthrough capacitor.

Calibration of the bridge of Fig. 10b follows the same procedure as before: a small collection of fixed carbon composition or metal film resistors with values from 10 Ω to 1000 Ω. With the value of differential capacitor that I was able to obtain, the bridge showed good nulls to 600 Ω, but above that value the dial marks were 'scrunched' too close together to be of any use (and the null was less sure).

Using the impedance bridge

One problem often encountered when using an impedance bridge in real situations is that they seem to work well only under the right circumstances. If the bridge is connected directly to the antenna feedpoint, it measures the actual impedance of the antenna. But when connected to the antenna through a random length of coaxial cable, the reflected impedance looking into the line will be related to, but different from, the actual

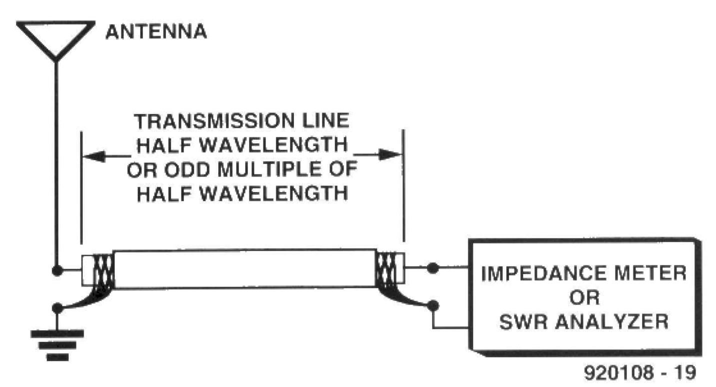

Fig. 11. Connect the impedance bridge to the antenna through a piece of transmission line that is an electrical half wavelength long.

antenna impedance as a function of line length. This problem is overcome by using a transmission line that is electrically one-half wavelength at the centre frequency of the band of interest (Fig. 11). This method works because the terminating impedance of a transmission line reflects itself every half wavelength down the line. This phenomenon was implied in Figs. 4b through 4d by the fact that the sharp voltage minima occur exactly half wavelength apart along the length of the line.

The physical length and electrical length of a transmission line are not the same. The electrical length is shorter than the physical length by the velocity factor (V). Calculate the physical length, L, required for half wavelength using the antenna length equation, and then multiply by V:

$$L = 150\ V / F \tag{6}$$

Where L is in metres and F is in megahertz. Typical values for V are 0.66 for polyethylene dielectric coax, 0.80 for polyfoam dielectric coax and 0.7 for Teflon® dielectric coax.
Once the antenna is adjusted, or the measurements are completed, you can substitute a random length of the same sort of coaxial cable for convenient run to the receiver or ham rig. The length of the cable does not affect the antenna adjustment (only radiator element length does that), but does affect the measurements.

Noise Bridges

A noise bridge is a special impedance bridge designed such that a noise generator circuit replaces the signal generator. An ideal gaussian noise generator creates a large amount random signal at all frequencies from DC to daylight, although for practical circuits it is only reasonable to expect operation from VLF through the upper end of the HF spectrum without some enhancements. Noise generators typically use shortwave receivers as the detector.

Figure 12 shows the circuit for a practical noise bridge. The bridge section is similar to others previously discussed. It can use the same trifilar transformer for Tr1 as the impedance bridge above. One difference between this bridge circuit and the other is that no hard-to-get differential capacitor is needed; an ordinary, but good quality, variable capacitor can be used. It is preferable to use a 'straight line capacitance' rather than 'straight line wavelength' model, as indicated by having the rotor shaft centred on the semicircular rotor plates.

Note that this circuit will also work as a straight impedance bridge if the noise source is replaced with a sinewave RF signal generator.

The noise generator circuit consists of a 6.8-V zener diode. All zeners operate in an avalanche mode that is inherently noisy. We use this phenomenon, which is ordinarily considered a defect, to make wideband pseudo-gaussian noise signals. The harmonic content, effect (and identifiability) of the noise source is enhanced by using a 1000-Hz pulse generator (a 555 timer, IC1) to chop the zener diode drive voltage at the 1000-Hz rate.

Two controls are provided in this circuit. The 'R' control provides a means for measuring the resistive component of impedance. This potentiometer should be a high quality carbon unit. The capacitor is an ordinary variable capacitor, and is used for the 'X' control (reactance). The 'X' control is adjusted to mid-scale when a 50-Ω resistive dummy load is connected to the ANTENNA port, and then BALANCE control C14 is adjusted for minimum noise.

Figure 13 shows how the noise generator is connected into the antenna circuit. Adjust the receiver frequency to the frequency in the centre of the desired band of operation. Turn on the noise generator. You should hear a noise 'hiss' of about S6 through S9. First, try connecting a 50-Ω dummy load to the antenna port. Adjust 'R' for a null in the noise. Then adjust 'X' to null. Note that there is some interaction between 'R' and 'X', so adjust both several times until the best

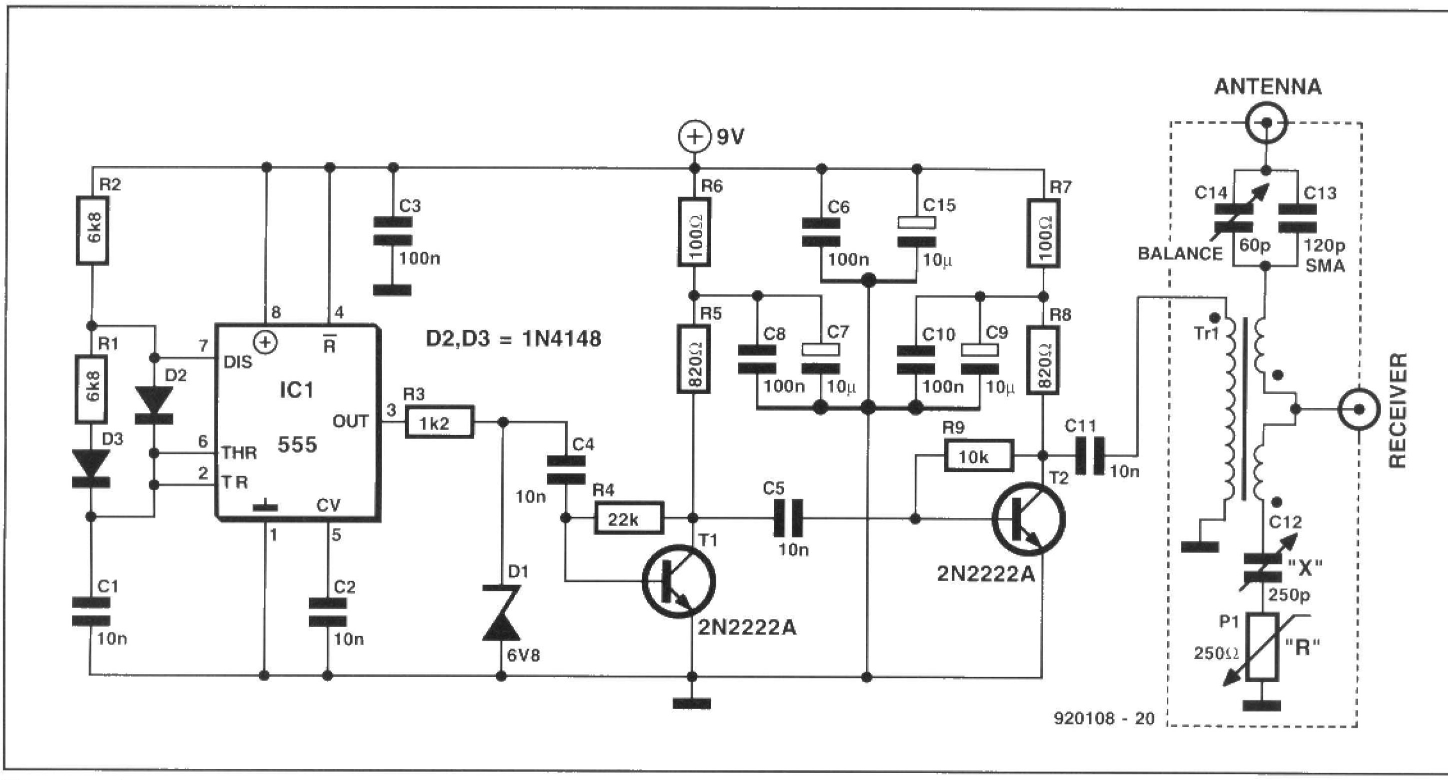

Fig. 12. Noise bridge circuit.

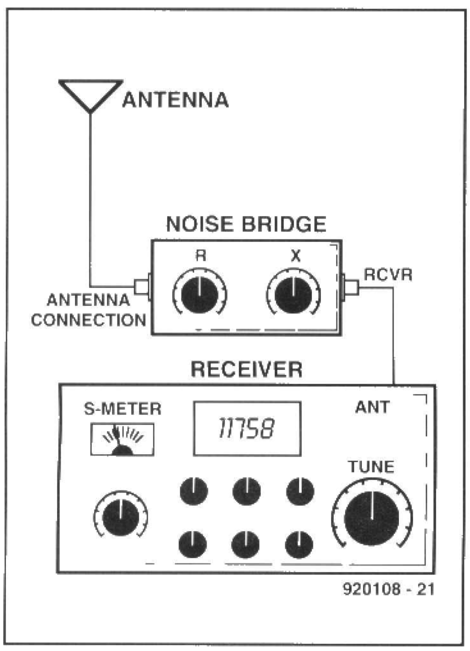

Fig. 13. Connecting the noise bridge to an antenna, using a shortwave receiver as the detector.

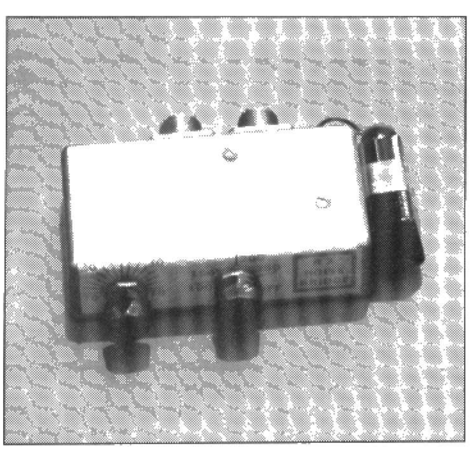

Fig. 14. Palomar Engineers noise bridge.

overall null is reached. The value of the 'R' control, in ohms, is the resistive component of the antenna impedance, while the value of the 'X' control provides a relative indication of the magnitude of the reactance. The 'X' can be calibrated for specific frequencies using standard value inductors and capacitors shunted by a 50-Ω resistance.

Once you get the hang of using the noise bridge with a dummy load, try it with a real receiver. Ham operators are advised to use a general coverage receiver for the detector, rather than a ham band only receiver, because the actual resonance point of an antenna might be out of the band (and therefore unhearable on a ham band only receiver).

Figure 14 shows the Palomar Engineers (Box 455, Escondido, CA, 92033, USA; Phone 619-747-3343) R-X Noise Bridge, a low-cost version of this instrument. I have used this one for several years and found it quite competent for the intended applications. Another low-cost instrument is the MFJ Enterprises, Inc. (Box 494, Mississippi State, MS, 39762, USA; Phone 601-323-5869).

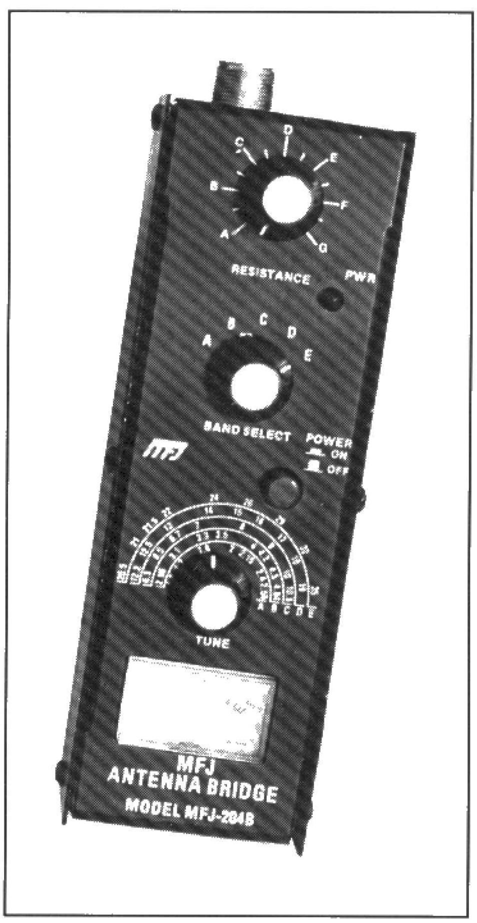

Fig. 15. Self-contained, handheld antenna bridge (MFJ).

Self-Contained VSWR Analyzers

Another MFJ product is the MFJ-204 self-contained VSWR meter shown in Fig. 15. This handheld instrument contains a bridge circuit, wideband amplifier and multi-band RF signal generator in one package. The resistance control is adjusted, with the signal generator tuned to the correct frequency, until a null is read on the meter. A conversion chart makes it possible to find the impedance corresponding to the resistance setting. More sophisticated instruments of the same sort include the MFJ-207 HF SWR Analyzer and the MFJ-208 VHF SWR Analyzer. These instruments do not have a resistance control, but use other methods to directly measure the VSWR and display it on a calibrated analogue meter. ■

80C32 COMPUTER APPLICATION NOTES - 2

Following the eight-instalment '8051/8032 assembler course' featured in last year's issues of *Elektor Electronics*, this column presents design ideas, programming examples and hardware experiments based on and around the popular 80C32 single-board computer. All descriptions are kept as brief as possible to ensure that a wide variety of subjects can be presented. Apart from the knowlegde you have hopefully gathered from the assembly language course, you will need the following to get going with the application notes: an 80C32 SBC (with extension board) running the EMON51 system monitor from EPROM, and a PC running the EASM51 assembler. Both programs are contained on the assembler course diskette, no. 1661. This month we have for you an infra-red command generator, and an intelligent windscreen wiper control.

RC5-code generator

Design by Dr. M. Ohsmann

INFRA-RED remote control of audio and video equipment has been the subject of several articles in *Elektor Electronics*. This 80C32 application note describes how the 80C32 single-board computer may be used to generate infra-red remote control commands to the RC-5 standard, as used for most Philips audio/video equipment. This application is based on the assumption that the RC-5 codes are transmitted to the equipment via wires. There are no functional differences with infra-red control as far as the commands and their logic structure are concerned; only a carrier is not used. Figure 1 recaps the structure of the 14-bit dataword.

The program shown in Fig. 3 is not available on disk, so you have to type it in yourself, and assemble it using EASM51. Next, the 80C32 computer (SBC) is connected to the PC as shown in Fig. 2, and the assembled code is conveyed to the SBC with the aid of V24COM. It is assumed that the SBC runs the monitor program, EMON51.

Start the code generator at address 4100H. The program will report with two question marks. You may enter either an A (for address) or a C (for command), followed by a decimal number. When an A is typed, the decimal number is the address to which all subsequent commands are sent. When a C is typed, the decimal number determines the command number transmitted after the carriage return (CR) key is pressed, whereupon the command is sent to the last defined address.

Let us assume that we want to transmit a 'pause' command to a CD player. This is achieved by typing A20 (return) C48 (return) without spaces. After entering a new address or a command, the program reports again with two question marks, prompting you to send further commands. The main program that takes care of this is short, and really serves as an example only to show how the subroutine RC5SND may be used.

RC5SND starts at line 38, and takes control of port line P1.0. The equipment address is read from register R1, and the command from register R0. In this way, RC5SND may be used as a module for all kinds of remote control applications based on the RC-5 code set. Evidently, the use of RC5SND is not limited to the 80C32 SBC — it may also be implemented on any other 8051-based microcontroller system. There is, however, one important condition: the system must run at a clock speed of 12 MHz, since

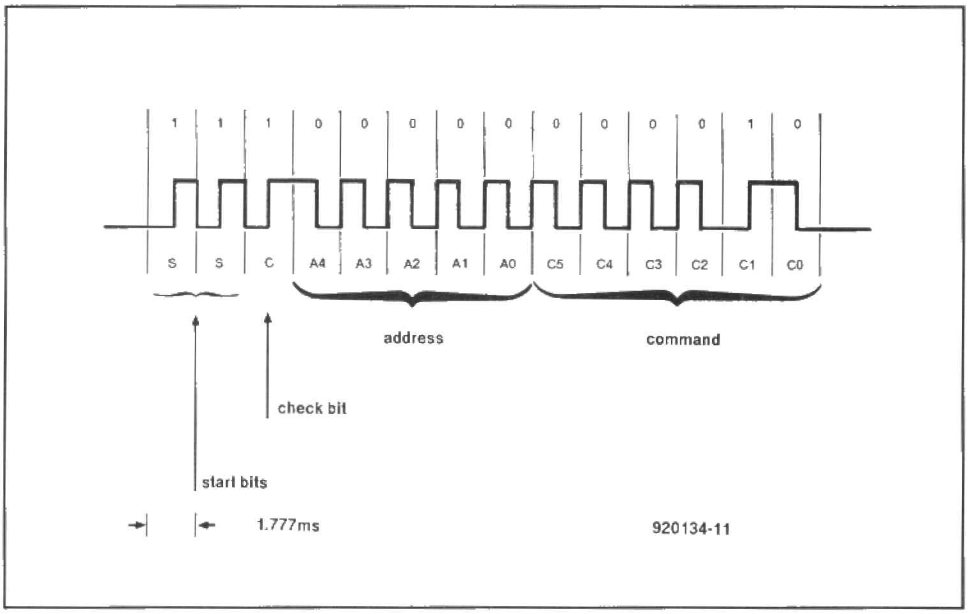

Fig. 1. RC-5 dataword structure: 14 bits are bi-phase encoded, and transmitted serially. The check bit toggles every time a command is transmitted.

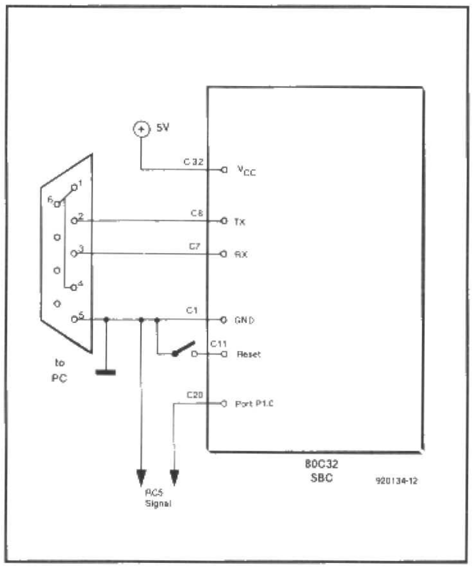

Fig. 2. These interconnections are required between the 80C32 SBC and the PC to enable the program to generate RC-5 codes.

```
; RC5 code generator for EMON51 + 80C32 SBC running at 12 MHz
;
P1      EQU  090H            ; special function register
ACC     EQU  0E0H
;
        ORG  050H            ; internal RAM, lowest free address
RC5cnt  DS   1               ; temporary counter for bit-number etc.
RC5adr  DS   1               ; RC5 equipment address stored here
Toggle  DS   1               ; toggling control bit
;
        ORG  4100H           ; program start address
        MOV  RC5adr,#20      ; CD Player as default RC5 equipment
        MOV  Toggle,#0       ; set toggle bit to 0
MAIN    MOV  DPTR,#TXT0      ; Prompt
        LCALL STXT
        LCALL GETU           ; Command "C" or "A"
        CJNE A,#'C',noCOM    ; not a "C"
        LCALL GETD           ; "C" entered, get RC5 command number
        MOV  R0,A            ; Command in R0
        MOV  R1,RC5adr       ; Address in R1
        LCALL RC5SND         ; transmit RC5 command
        LJMP MAIN            ; and start again
noCOM   CJNE A,#'A',MAIN     ; not an "A" so start again
        LCALL GETD           ; "A" entered, which address?
        MOV  RC5adr,A        ; store for later
        LJMP MAIN            ; and start again
;
TXT0    DB 13,10,'??',0
;
RC5SND  EQU  $             ; send RC5 train, R1:Address, R0:Command
        MOV  A,Toggle      ; invert toggle bit
        XRL  A,#00100000B
        MOV  Toggle,A
        MOV  A,R1          ; Address bits
        ORL  A,#11100000B ; two start bits = 1 ; Control bit=1
        ORL  A,Toggle      ; control bit = toggle bit ++++++
        MOV  R1,A          ; gives most significant byte
        MOV  A,R0          ; 6 command bts
        RL   A             ; shifted to the left two times
        RL   A
        MOV  R0,A          ; gives least significant byte
        MOV  RC5cnt,#14    ; send 14 bits
RC511   MOV  A,R0          ; shift 16-bit shift register
        RLC  A             ; Time counts to 2nd half of last bit
        MOV  R0,A
        MOV  A,R1
        RLC  A
        MOV  R1,A
        CPL  C             ; invert highest bit
        MOV  P1.0,C        ; gives first half bit of RC5 train
        MOV  A,#220        ; takes so long (approx.  )
        LCALL TIME
        CPL  P1.0          ; invert (biphase encoding)
        MOV  A,#218        ; 2nd half of bit
        LCALL TIME         ; wait
        DJNZ RC5cnt,RC511 ; and continue until all 14 bits sent
        CLR  P1.0          ; pause (a bit early)
        MOV  RC5cnt,#5     ; approx. 5*200*4 microsec.
RC512   MOV  A,#200
        LCALL TIME
        DJNZ RC5cnt,RC512
        RET                ; that's all
;
TIME    NOP                ; 4*RC5tim   microseconds
        NOP                ;  +4        microseconds for RET and CALL
        DJNZ ACC,TIME
        RET
;
ccGETU  EQU  014H   ; EMON51 command to get upper case characters
ccGET10 EQU  013H   ; EMON51 command to get decimal number
ccSTXT  EQU    2H   ; EMON51 command to send text
COMMAND EQU  030H   ; EMON51 command RAM location
MON     EQU  0200H  ; EMON51 entry address
;
STXT    MOV COMMAND,#ccSTXT   ; send text from DPTR
        LJMP MON
;
GETD    MOV COMMAND,#ccGET10  ; fetch decimal number via V24
        LCALL MON
        MOV  A,@R0            ; put it into accumulator
        RET
;
GETU    MOV COMMAND,#ccGETU   ; get upper case characters
        LJMP MON
        END
```

Fig. 3. Listing of the RC-5 code generator program.

that determines the timing of the RC-5 pulse train.

The line marked with '++++++' is included to make sure that the check bit toggles (i.e., changes from 0 to 1 or from 1 to 0) at each transmission. The check bit control is copied from 'classic' RC-5 code generator ICs like the SAA3006, when a key is released on the remote control. The RC-5 receiver takes the change of the check bit to mean that a new command follows. If you wish to simulate command repeating (i.e., a key kept pressed), this line must be omitted from the listing.

For further reading:

1. Universal RC-5 code infra-red receiver. *Elektor Electronics* January 1992.
2. SAA3006 data-sheets, Philips Semiconductors.
3. Infra-red receiver for 80C32 single-board computer. *Elektor Electronics* April 1993.

Windscreen wiper control

Design by Dr. U. Kunz

THIS is a stand-alone application of a stripped-down version of the 80C32 single-board computer. The cost of making a BASIC version of this computer can be reduced significantly by using an 80C32 controller in combination with an external EPROM that contains the BASIC interpreter copied from a 8052AH-BASIC. Furthermore, quite a few parts may be omitted from the board, so that the final wiper control will be hardly more expensive than a conventional, i.e., non-microprocessor controlled, circuit, which it beats hands down as far as user friendly operation is concerned. Even if you are not interested in the present application, it is still worth while to have a look at the proposed program, which illustrates the use of interrupts (which are always a bit of a problem with the 8052 BASIC interpreter).

Figure 4 shows you how to tackle **the hardware**. The additional elements are two press-keys, a reset switch and a relay driver. The on/off switch shown in the left-hand corner of the diagram is optional, because the controller board may be powered by the car battery via the ignition switch. The connections to make in the car are straightforward: supply power (12 V), and a parallel connection of the relay contacts to the existing wiper switch. In addition, the two press-keys must be fitted at an easily accessible location.

The shaded sections of the circuit diagram indicate parts that are not required for the present application. An example is the EPROM programming circuitry. The parts list printed here reflects the omissions, giving only those parts that are absolutely necessary.

The **software** is based on the assumption that the 80C32 runs in BASIC, and EPROM IC7 contains the program listed in Fig. 5. When the system is switched on via S1, line 30 sets port 1 to 128, so that bit 7 is taken high, and the relay, Re, is energized via transistor T1. The switch-on time depends on the value of variable 'S'. The indicated value (1000) results in an on-time of about 1 second. Instruction PORT1 = 1 in line 70 causes bit 7 of port 1 to return to logic low, so that the relay is switched off. If the relay contacts are correctly wired in parallel with the contacts of the wiper switch on the dashboard, the windscreen wipers do one wipe. The time set by 'S' for one wipe action is only an indication. Other values than the one given may be required for particular cars, allowing for the speed of the windscreen wiper motor.

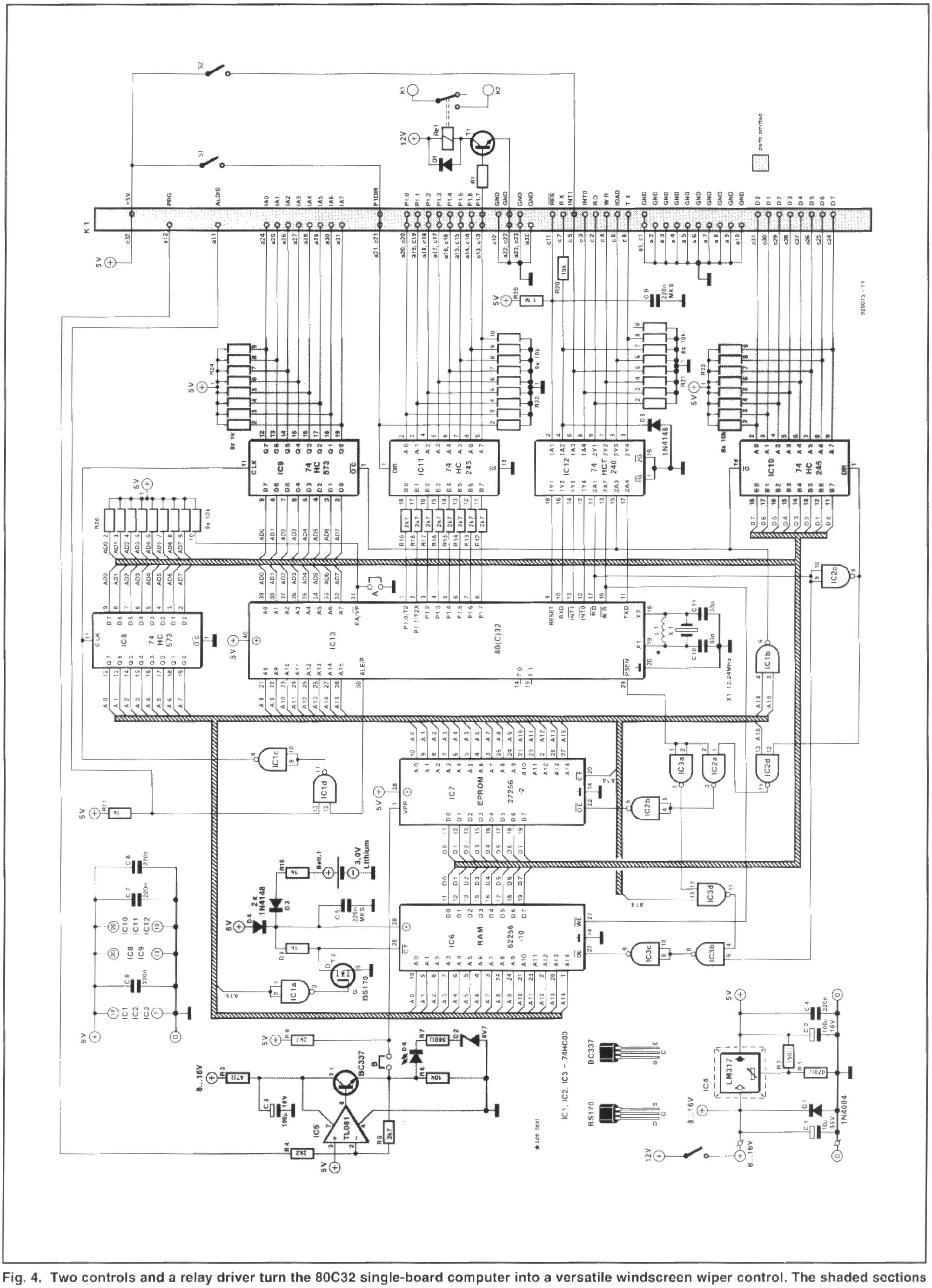

Fig. 4. Two controls and a relay driver turn the 80C32 single-board computer into a versatile windscreen wiper control. The shaded sections indicate components that may be omitted from the 80C32 SBC.

COMPONENTS LIST

Resistors:

1	470Ω	R1
1	150Ω	R2
2	1kΩ	R9;R11
8	2kΩ7	R12-R19
1	15kΩ	R20
3	9-way 10kΩ SIL	R21;R22;R26
1	1MΩ	R25

Capacitors:

1	10µF 35V radial	C1
1	100µF 16V radial	C2
6	220nF	C4-C9
2	33pF ceramic	C10;C11

Semiconductors:

1	BS170	T2
1	1N4004	D1
1	1N4148	D5
3	74HC00	IC1-IC3
1	LM317	IC4
1	62256-LP2	IC6
1	27C256-2	IC7
1	74HC573	IC8
1	74HC245	IC11
1	74HCT240	IC12
1	80C32 or 8052AH-BASIC	IC13

Miscellaneous:

2	Push-to-make press-key	S1;S2
1	On-off switch (optional)	
1	12-V relay with make contact	Re
1	12MHz or 15MHz quartz crystal	X1
1	Lithium battery with PCB holder	
1	Heatsink SK129/25,4	
1	Printed circuit board 910042	

At the same time, a counter 'Z' is started in line 90. More wiper action may be required to keep a clear view through the windscreen. This is achieved by pressing key S2. The state of this key is continuously monitored by line 110. If pressed, S2 causes port 1 to go low since it determines the data direction, and the inputs of port 1 are held low by resistor array R22.

Next, counter 'Z' is halted, and the program continues from line 120, where the relay is energized again for a time defined by 'S'. Another counter, 'W', is started in line 170. The state of 'W' is continuously compared to that of 'Z' (line 210), and an overflow results in a wiper action. After 'W' has been reset in line 270, the program jumps to line 160, where a new wipe cycle is started.

In practice, the software results in the following operation of the wiper control. After pressing S2, the wiper operates at constant intervals. However, since the amount of rain on the windscreen is hardly ever constant, it will be required after a while to define a new interval. This is achieved by pressing S1. This press-key is connected to the microcontroller's interrupt input, whose state is continuously checked in line 180. When S1 is pressed, the program enables an interrupt for the next cycle, and then branches to the start of line 10. Line 20 resets the counter states to 0, and a new wipe action is started from line 30 onwards. Next, counter 'Z' is increased until S2 is pressed, which defines the new wiper interval time.

The use of only two keys to control the wiper interval time affords a comfortable way of wiping your way through all types of downfall, from a heavy downpour to a slight drizzle, without ever being blinded by water on the windscreen, or hearing the scraping sound of wipers on a dry windscreen.

Copying the BASIC interpreter

Many readers still seem to be unable to transfer the BASIC interpreter contained in the 8052AH-BASIC into an EPROM. This interpreter is required for the present application, which is based on an 80C32 running it from an EPROM type 27C64. So, a few words on how it is done.

The copy is made with the aid of the 8052AH-BASIC computer described in Ref. 1. Fit a 27C64 EPROM in socket IC6, apply a programming voltage of 12.5 V, and run the copy program described in Ref. 2. Enter the following parameters:

- Starting data address: 8192;
- Ending data address: 16384;
- Starting PROM address: 32768.

When the programming is successful, the EPROM will contain the complete interpreter from address 0 onwards. Bend EPROM pins 26 and 27 sideways, and connect them to pin 28. Bend pin 22 sideways, and solder it to an approximately 8-cm long piece of wire. Next, the EPROM is soldered 'piggy-back' on a 32-Kbyte RAM Type 62256. All EPROM pins except 22, 26 and 27 are soldered straight to the corresponding RAM pins. Insert this 'double-IC' into position IC6 on the computer board, and connect the wire to pin 1 of IC3. This leaves position IC7 free for EPROMs with BASIC programs. Populated in this way, the 80C32 single-board computer mimics a 8052AH-BASIC system, although it lacks an EPROM programming facility.

References:

1. BASIC computer. *Elektor Electronics* November 1987.
2. 8032/8052 single board computer.

```
10   REM wiper interval con-
     trol
20   Z=0 : W=0 : S=1000
30   PORT1=128
40    FOR I=0 TO S
50    NEXT I
60    PRINT "WIPE!"
70   PORT1=1
80    DO
90   Z=Z+1
100   PRINT Z
110   UNTIL PORT1=0
120  PORT1=128
130   FOR I=0 TO S
140   NEXT I
150  PORT1=1
160   DO
170  W=W+1
180   ONEX1 10
190   CLEAR I
200   PRINT Z,W
210   UNTIL W>Z
220  PORT1=128
230   FOR I=0 TO S
240   NEXT I
250   PRINT "WIPE!"
260  PORT1=1
270  W=0
280   GOTO 160
290   RETI
300   END
```

Fig. 5. 8052 BASIC listing of the windscreen wiper control program.

Elektor Electronics May 1991.

Next time:
X2404 EEPROM interfacing to an 8051.

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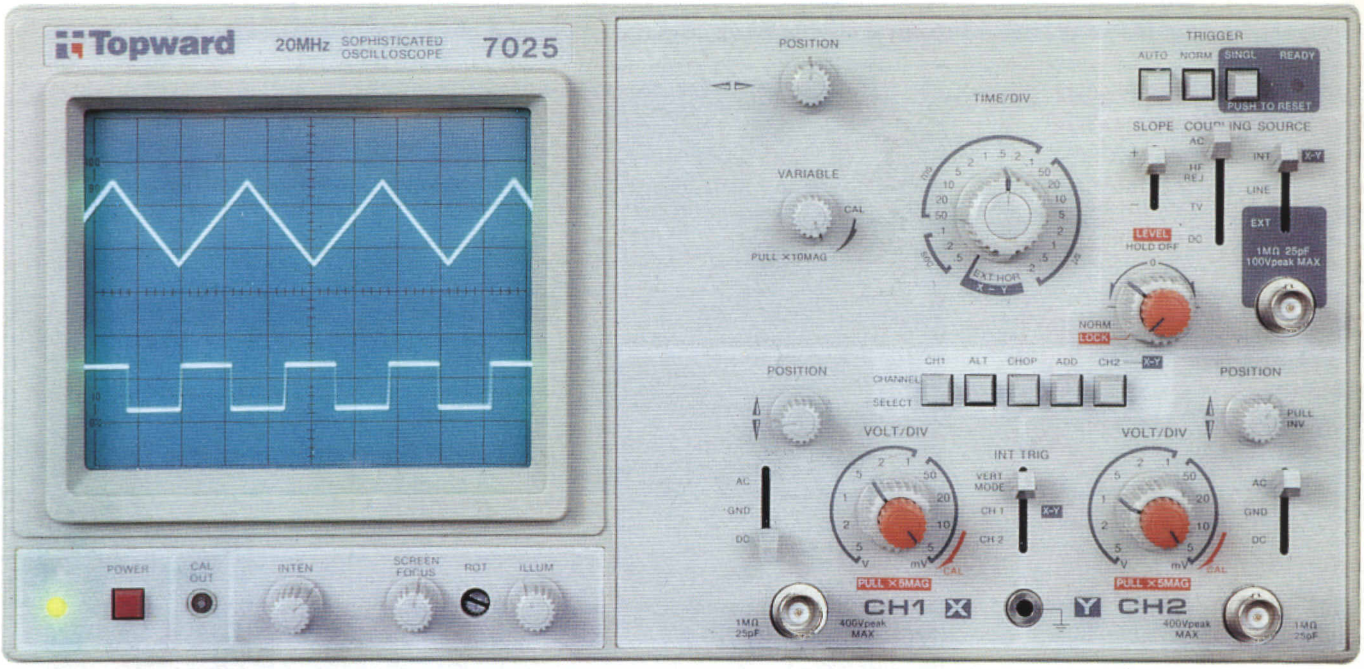

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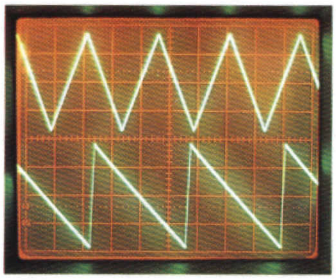

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